RIDER'S VOLUME XV

HOW IT WORKS

by
The Technical Staff
of
John F. Rider Publisher, Inc.

PREFACE

One of the characteristics of the radio industry is the constant improvement in the design of receivers which advances their performance. This has been most noticeable in the receivers which have been marketed since the conclusion of the war. There seems to be little doubt that the vast amount of research that was carried on during the period of conflict is reflected to a certain extent in much of the present-day receiving equipment.

From the point of view of the man who traces out the troubles in a set and repairs them, innovations in electrical and mechanical design present a two-fold problem. Not only must he learn the electronic theory underlying the new features and their manner of functioning, but he must acquaint himself with the mechanical construction and operation of numerous other new gadgets. He must recognize the fact that an apparently forgotten unsoldered end of wire nestling beside a coil is not an oversight on the part of someone in the manufacturer's assembly line, but a capacitor; he must learn that capacitors are concealed within the base of some i-f transformers and that they are not only trimmers but fixed capacitances which have plates that are in common with the trimmers and why they are there, and so on ad infinitum.

It is the purpose of this book to present to the users of Rider's Volume XV some of the ingenious brain children of American radio receiver design engineers. When it was felt that some theoretical background was necessary for a better understanding of the problems and their solutions, the fundamentals of the subject have been briefly covered and these basic matters applied to the new problems. In practically every instance receivers covered in Rider's Volume XV have been used as illustrations, either electrical or mechanical; in those instances where other circuits have been considered, they were chosen because it was thought that they were trends with which the radio repairman should be familiar.

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ALIGNING WITHOUT MANUFACTURER'S DATA

This section will explain how a conventional a-m superheterodyne receiver, for which no manufacturer's alignment data are available, can be aligned. The first step is to obtain the necessary data by means of various measurements on the receiver. Once these data are obtained, the receiver is aligned in the conventional manner. For purposes of illustration, it will be assumed that no circuit schematic is available.

The data required for alignment consist of the following:

1. Locations and adjustment frequency of i-f trimmers.
2. Location and adjustment frequency of wave trap.
3. Locations and adjustment frequencies of the oscillator shunt trimmers for the various bands.
4. Locations and adjustment frequencies of the oscillator series padders for the various bands.
5. Locations and adjustment frequencies of the mixer and r-f trimmers for the various bands.

Usually, the amplifier trimmer capacitors are mounted in the i-f coil cans. Where iron-core trimmer coil slugs are used, they will be found in their respective coils. In straight a-m receivers (no f-m), where the i-f transformer includes a trimmer capacitor (generally having a slotted screw head projecting through the can top) and a trimmer coil slug (generally having a slotted brass screw projecting through and centrally located on the can top), the coil slug will most commonly be the alignment adjustment for the transformer primary while the trimmer capacitor will be the transformer secondary alignment adjustment. This should be checked by resistance measurements.

Determining Unknown I-F Peaks

One of several methods of determining the unknown i-f peak is to connect the i-f section of a signal tracer between the receiver mixer plate and circuit ground. This, in the case of ac-dc receivers, need not be the chassis of the receiver. The signal tracer is then tuned for a peak starting at 550-kc and decreasing to 100-kc. These frequencies are given as limits since the i-f peak for the broadcast band will be somewhere between these two frequencies. The frequency at which a peak is reached on the signal tracer i-f tuning eye will be the i-f peak (approximately) since the beat between the incoming broadcast signal and the receiver local oscillator produces a difference frequency equal approximately to the i-f peak. To be certain that the signal tracer is tuned to the i-f peak and not to the second harmonic of the beat frequency, the signal tracer should be tuned to one-half the frequency at which the peak was found; if no peak is obtained at this point, then the first frequency is the i-f peak. As a further check, it is suggested that the signal tracer be tuned to the second harmonic of the beat frequency, which will be found at twice the frequency of the i-f peak.

The above explanation will be simplified by the following example. Suppose that beginning with the signal tracer at 550 kc, the frequency is decreased slowly so that when 456 kc is reached the tuning eye of the signal tracer closes. The signal tracer is then set at half 456 kc or 228 kc and at this setting there should be no response in the signal tracer. With the signal tracer set at twice 456 kc or 912 kc, the tuning eye should close. If the above conditions are fulfilled, then the *approximate* i-f peak is 456 kc.

As a final check, the receiver should be tuned to each succeeding broadcast station down to the low-frequency end of the dial, and the signal tracer should indicate a peak each time a station is tuned in.

A second method of finding an unknown intermediate frequency is very similar. The procedure is to connect the signal generator (whose output is modulated) to the grid of the last i-f tube and to short out the oscillator section of the tuning condenser. Starting at 550 kc, the signal-generator frequency is decreased until the output meter shows a reading. The speaker is preferably left connected while this operation is carried out, so that the audio note will be heard as the peak is passed. The frequency at which the note is first heard will

probably be the intermediate frequency. To determine that this is not a sub-harmonic of the actual intermediate frequency, the signal generator should then be set to twice the frequency at which the note was first heard; if no signal is heard at this point, then the first frequency is the i-f peak. As a further check, it is suggested that the signal generator be set to half the i-f and the signal should then be heard.

A concrete example will simplify the above explanation. Beginning with the signal generator set at 550 kc, the frequency is decreased slowly and when 456 kc is reached, a signal is heard in the speaker. The signal generator is set at twice 456 kc or 912 kc and at this setting no signal is heard. With the signal generator set at one-half 456 kc or 228 kc, a signal is heard but with less strength than the true i-f peak signal. If the above conditions are fulfilled, then the *approximate* i-f peak is 456 kc.

A third method of obtaining the i-f peak follows: the signal generator is connected to the grid of the last i-f tube and is set approximately at 550 kc. The receiver oscillator must be shorted out. The last i-f trimmer capacitor is screwed tight. The generator frequency is decreased until an audio peak is heard in the speaker. The frequency of the generator will now be the lower limit of the i-f trimmer tuning range. This value is recorded. The same trimmer capacitor is then loosened until it is almost all the way out. The frequency of the generator is increased slowly to obtain a second audio peak, the generator frequency at this point will be the upper limit of the i-f trimmer tuning range. This value is also recorded. These two frequency limits, added together and divided by two, will give the approximate frequency of the i-f amplifier transformer bandpass.

As an example of this method, if the frequency of the generator producing an audio peak when the trimmer is at maximum capacitance (plates tightened) is 410 kc and the frequency of the generator producing an audio peak with the trimmer at minimum capacitance (plates loose) is 500 kc then the i-f peak is $\frac{410 + 500}{2}$ or 455-kc.

A more accurate method of determining the i-f frequency is; the signal tracer is connected from the oscillator anode or grid to ground. The receiver tuning capacitor is set at approximately half mesh and is not disturbed until the intermediate frequency is determined. The oscillator frequency is determined with the oscillator shunt trimmer at minimum and at maximum capacitance. The signal tracer is set to the average of the frequency values found and the oscillator shunt trimmer is adjusted until maximum response is obtained on the signal tracer.

As an example of this step, if the oscillator frequency is 1260 kc when the oscillator shunt trimmer is at maximum capacitance, and the oscillator frequency is 1490 kc when the oscillator shunt trimmer is at minimum capacitance, then the oscillator shunt trimmer is adjusted to the average frequency of 1375 kc.

Determining Frequency of R-F Circuits

Since the oscillator frequency is now known, all that is necessary to derive the intermediate frequency is to determine the frequency to which the r-f circuits are tuned. This may be done as follows:

The i-f transformer is disconnected from the converter plate and a 10,000-ohm (approximately) resistor is connected between the converter plate and B+; the signal tracer i-f section is connected between the converter plate and chassis. A signal generator is connected to the receiver input terminals, and is set to approximately the frequency to which it is thought the receiver i-f circuits are tuned, the signal tracer being tuned for maximum response and this reading is recorded. A series of adjustments of the generator and tracer is then made centering about the probable frequency to which the r-f circuits are tuned, with the signal-tracer sensitivity level being recorded for each adjustment. The frequency at which the signal tracer shows greatest response is the frequency to which the receiver r-f circuits are tuned.

This frequency value when subtracted from the value of the oscillator frequency (previously determined) is equal to the difference frequency at the plate of the converter tube. The i-f amplifier may be aligned at this value.

The table below is an example of readings on the signal tracer to secure the frequency of the r-f circuits.

Signal Tracer Frequency (kc.)	Signal Tracer Sensitivity Level
1050	2½
1000	4
950	10
935	15
925	20
915	22
900	15
875	10

It will be noticed that the highest response on the signal tracer was secured at 915 kc, so that this is the frequency to which the signal circuits are tuned. Since the oscillator frequency was found to be 1375 kc (above), the difference frequency between the oscillator and signal frequencies equals 460 kc and this may be considered the intermediate frequency of the receiver in this example.

The i-f transformer is then reconnected to the converter plate and the 10,000-ohm resistor and the signal tracer removed. The signal generator and output meter are then connected for conventional i-f alignment and the i-f trimmers are aligned at the determined frequency.

If the manufacturer had set the i.f. of this receiver at 450 kc (which we cannot ascertain), ordinarily it would not make the slightest difference if we align the i.f. at 460 kc, since the i-f and oscillator circuits are designed to have considerable tolerance. The selectivity and sensitivity of the receiver will not be decreased if the receiver is aligned at the frequency determined by the above method.

Locating Trimmers

The following information will aid in locating oscillator and r-f trimmers. Where different capacity sections are used in a tuning capacitor, the section having the smaller plates, a fewer number of plates, or differently shaped plates, will probably be the oscillator tuning capacitor since the oscillator in almost all sets operates above the signal circuits. The oscillator and r-f shunt trimmers will often be found on the tuning capacitor frame near their respective sections. Oscillator and r-f trimmers for short-wave bands will be located physically close to their respective coils. To determine which coils are in the circuit for a particular band position, without laborious tracing of circuits, measure the voltage across the coils with an r-f voltmeter. Assuming the receiver is operating properly, those coils which are in the particular short-wave circuit which is operating will have voltages impressed across them, while those not in the circuits, (open or shorted out) will not have voltages across their terminals. In receivers incorporating an f-m band, the f-m oscillator and r-f trimmers will be found on or near their respective f-m tuning capacitors. In older receivers making use of plug-in coils, the trimmer will be located in the respective coils or beneath the respective coil sockets.

The oscillator and r-f shunt trimmers should be aligned at a frequency slightly below the high end of the band while the oscillator series padder should be aligned at a frequency slightly above the low-frequency end of the band. Thus the shunt trimmers would be aligned with the tuning capacitor plates just meshed and the padder would be aligned with the tuning capacitor plates not quite fully meshed; the tuning capacitor must be rocked* during the padder adjustment. For receivers having the conventional range on the standard broadcast band of 550 kc to 1500 kc, the high alignment frequency for both oscillator and r-f circuits would then be 1400 kc and the oscillator low alignment frequency would be 600 kc. In those receivers having extended ranges at the high-frequency end of the band, the alignment would be effected as follows: for those receivers having a band range to 1600 kc, align the oscillator shunt trimmer at 1500 kc and the r-f trimmers at 1400 kc; for those receivers having a band range to 1700 kc, align the oscillator shunt trimmer at 1700 kc and the r-f trimmers at 1600 kc; for those receivers having a band range to 1800 kc, align the oscillator shunt trimmer at 1800 kc and the r-f trimmers at 1700 kc. These alignment frequencies should give maximum sensitivity and best tracking over the band on the respective ranges. Those receivers having a slightly extended range at the low end of the band to 540 or 535 kc, may be aligned at 580 kc instead of 600 kc.

Correcting Dial Calibration

The dial calibration may be corrected by tuning in a known broadcast station somewhere

*For an explanation of rocking, see the last section of this article.

in the middle of the broadcast range and setting the tuning dial pointer to coincide with the frequency on the dial for this station.

Assuming the worst possible condition, where the receiver dial is missing, the alignment of the short-wave bands would first require the determination of the band ranges. An output meter is connected at the audio output of the receiver and the tuning capacitor of the receiver is opened completely. Feed a modulated signal from the signal generator into the receiver input terminals, and tune the generator through the frequencies to which this particular short-wave band may be tuned starting from the *highest* possible frequency. When the generator signal is heard in the speaker, it may be an image signal. It cannot be a signal-generator harmonic if the generator is reduced from the *highest possible* frequency. It is necessary then to connect the oscillator section of the signal tracer from the receiver oscillator anode or grid, to ground and tune the signal tracer for peak response at a value which is equal to the signal-generator frequency *plus* the i.f. and repeat this operation at a value equal to the signal-generator frequency *minus* the intermediate frequency.

Thus, for example, if a signal is heard in the speaker when the generator is set at 18 mc and the receiver i.f. is 455 kc, then the signal tracer should be tuned about 18.455 mc and 17.545 mc. Using these frequencies as examples, if the signal tracer shows a peak at 18.455 mc and none at 17.545 mc, then the oscillator appears to be operating at a frequency *above* the r-f input signal frequency by a value equal to the i.f. If the signal tracer responds at 17.545 mc and does not respond at 18.455 mc, then the oscillator appears to be operating *below* the frequency to which the r-f circuits are tuned. The reason the oscillator is said to "appear to operate" above or below the signal frequency is because the short-wave oscillator shunt trimmer may be misaligned seriously and thus cause the oscillator to operate above or below just at this point on the tuning range.

Checking Oscillator Frequency

To check definitely whether the oscillator is operating above or below the signal frequency, the following procedure is suggested: the generator is reduced one megacycle in frequency from the first setting which should be the highest frequency of this particular band, and the receiver is tuned for maximum output indication on the output meter or in the speaker. The signal tracer is then tuned for maximum and the frequencies of the generator and tracer are recorded. The generator is then reduced one megacycle in frequency and the same process repeated. This is continued until the receiver tuning capacitor is meshed completely.

The recorded figures must then be analyzed carefully since they are essentially a calibration of signal-generator and receiver tuning versus oscillator frequency. If the figures show that the oscillator is consistently above the receiver tuning (signal-generator frequency) by approximately the i-f value, then the oscillator is operating *above* the r-f frequency. If the figures show that the oscillator is consistently below the receiver tuning (signal-generator frequency) by approximately the i-f value, then the oscillator is operating *below* the r-f frequency. Either of these two conditions, if consistent, indicate that the oscillator shunt trimmer is set at or near the correct point of operation. If, however, the recorded figures show the oscillator frequency to have no consistent relation to the r-f frequency, then it must be assumed that the oscillator shunt trimmer is aligned to the image frequency. If this latter condition exists, then the oscillator trimmer must be peaked to the proper frequency.

The receiver tuning capacitor should be turned almost to the completely unmeshed position and the signal generator set at the frequency previously found to be the approximate high end of the band. The generator is then tuned for maximum output from the speaker. The oscillator shunt trimmer is then tightened slowly (increasing capacitance) until a peak signal is obtained in the speaker. If no peak is obtained and the trimmer becomes tight, it should be loosened until the signal peak is obtained (other than the one originally obtained.) The previously described process of setting the generator and receiver to frequencies one megacycle below the previous setting, adjusting the signal tracer for peak response, and recording the generator and tracer frequencies, is repeated. The oscillator frequencies should now have a consistent relation to the r.f. being either above or below by the value of the intermediate frequency. If this condition now holds, the oscillator series padder

may now be adjusted by setting the receiver tuning capacitor. The oscillator shunt trimmer is then readjusted at the frequency at the high end of the band and the r-f trimmer is also adjusted for maximum output at this frequency.

Some receivers do not use a separate oscillator circuit for the short-wave band but depend on the second harmonic of the oscillator for the heterodyning signal to produce the receiver intermediate frequency. This should always be kept in mind when checking the oscillator frequency.

Determining Wave-Trap Frequency

The frequency to which a series wave trap is tuned may be found by the following method. With the receiver not operating, the high end of the wave-trap circuit is disconnected from the associated receiver circuits. A signal generator is connected across the wave trap (coil and trimmer capacitor) and an a-c vacuum-tube voltmeter is connected across the trimmer capacitor. The setup is illustrated in Fig. 1.

FIG. 1.—Signal-generator connections for determining range of wave trap.
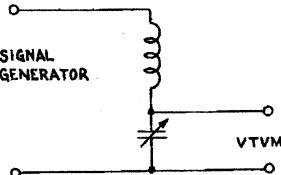
The generator is set approximately to the frequency at which it is believed the wave trap is tuned and the generator frequency is varied until the voltmeter responds. Wave traps are tuned to interfering signals so that most often the wave trap will be tuned to a frequency at or near the i-f peak of the receiver. It may be peaked, however, to the frequency of a local station which might overide all other stations received.

Once the approximate frequency to which the wave trap is tuned is found, the trimmer is measured at maximum capacitance and at minimum capacitance and the frequencies to which the wave trap could possibly be tuned are computed. The wave trap is then reconnected in the receiver, the generator and voltmeter removed and the wave trap is aligned for *minimum* output response to the frequency (within the wave-trap tuning range) at which most interference is encountered. It is preferable to align a wave trap following the alignment of the i-f amplifier and preceding the alignment of the oscillator and r-f circuits so as not to detune the r-f circuits by the adjustment of the wave trap.

High-Fidelity Alignment

The alignment of a high-fidelity receiver where no data are available, is more complicated than for a conventional receiver. The i-f amplifiers in high-fidelity receivers usually have a wider bandpass, which is obtained by overcoupling or detuning one or more of the i-f stages to obtain a double-peaked characteristic. Proper alignment of such high-fidelity receivers, therefore, requires that receiver fidelity characteristics be understood.

The fidelity of a radio receiver indicates the degree to which it accurately reproduces in the speaker, the modulation of the signal entering the receiver. The modulation of a radio-frequency carrier by an audio-frequency signal is really a heterodyning action and, therefore, results in beat frequencies corresponding to the sum and difference frequencies of the r-f carrier and the various modulating audio frequencies. For each audio frequency used to modulate the carrier, two radio-frequency signals are created, one equal to the carrier frequency *plus* the audio frequency and one equal to the carrier frequency *minus* the audio frequency. A modulated signal is then constituted of many side frequencies in addition to the original r-f carrier or center frequency and thus a modulated signal occupies a frequency band equal to twice the highest modulating frequency.

Exact reproduction of the modulation carried by a broadcast signal requires that the receiver bandwidth be great enough to pass the highest modulating frequency transmitted. In addition, high fidelity demands that the various frequencies constituting the modulated signal pass through the receiver with no change in their relative amplitudes. To illustrate this point, if a station broadcasts a program of symphony music, the transmitted signal may contain sideband frequencies up to 5000 cps (corresponding to notes emitted by such musical instruments as the clarinet, flute, violin, etc.) Faithful reproduction of the modulating frequencies contained in this transmitted signal requires that the receiver be capable

of amplifying equally all frequencies in the band from 5000 cps above the carrier frequency to 5000 cps below the carrier frequency. Uniform amplification of all frequencies in the modulated wave, which are within the i-f bandpass of the receiver, requires that alignment of a high-fidelity receiver be optimum in terms of the receiver design.

High Fidelity Bandpass Response

A high-fidelity receiver differs from a normal- or high-selectivity receiver mainly in the bandpass response of the i-f amplifiers, since the r-f signal circuits normally do not have enough over-all selectivity to cut the modulation sidebands to any degree. High-fidelity receivers employ i-f transformers designed to have resonance curves which are wide at the top instead of sharply peaked, so that sidebands will not be suppressed in the i-f amplifier to such a degree that the receiver will lack "highs" (high audio frequencies).

An i-f transformer sharply tuned as a result of loose coupling of the primary and secondary coils of the transformer has a resonance curve similar to the curve "A" shown in Fig. 2. With
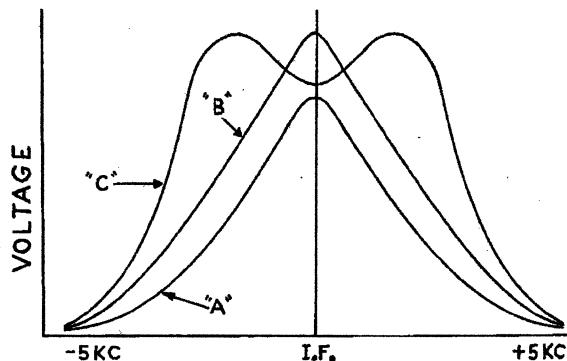
Fig. 2.—Variably coupled r-f transformer resonance characteristics.

the primary and secondary coils of the transformer placed closer to each other, at the point of critical coupling, the response curve would be similar to curve "B" shown in Fig. 2. When the transformer coils are placed even closer together, overcoupling takes place, and the response curve takes the form of curve "C" in Fig. 2.

Band-pass action in high-fidelity receivers is most often obtained by combining the double-humped response (curve "C") of an over-coupled transformer with the resonance curves (curve "B") of single-peaked transformers. The resulting over-all response curve of the i-f amplifier, when the circuits are so designed that the response is the same at resonance (center i-f) and at the two side peaks, is shown in Fig. 3.
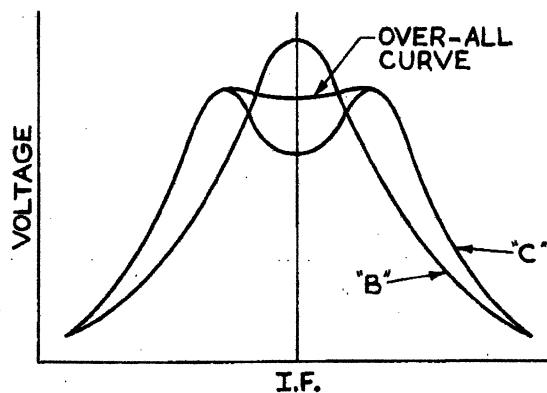
Fig. 3.—Over-all i-f amplifier response of high-fidelity receiver.

Receiver high-fidelity response, as shown by curve "C" in Fig. 2, is obtained mainly in conventional superheterodyne receivers by one of five methods. The most practicable and common method which is applied to one or more of the i-f transformers is the physical placement of the primary and secondary coils with respect to each other to produce overcoupling. The other four methods of obtaining high-fidelity band-pass response are: loading of the primary or secondary coil or both coils of the i-f transformers with a resistor thus lowering the Q of the circuit and decreasing the selectivity of the circuit; detuning of primary and secondary coils several kilocycles in opposite directions from the center intermediate frequency; staggering of the transformers, so that one transformer will be tuned several kilocycles above and one transformer will be tuned several kilocycles below, the center intermediate frequency to which a third transformer will be tuned; and use of triple-tuned transformers, having three resonant circuits with individual trimmers. These methods, with the exception of staggering and resistance loading, result in overcoupling of the coils on those transformers on which the above methods are used.

The intermediate frequency can be determined as explained under the procedure given at the beginning of this article. The single-peaked sharply tuned i-f transformers will, of course, be aligned at the i-f peak.

Aligning Triple-tuned I-F Transformers

Many high-fidelity receivers using triple-tuned i-f transformers may be aligned as follows: with the output meter connected across the receiver audio output, the signal generator, set to the intermediate frequency, is connected to the grid of the last i-f tube. If the last i-f transformer is single-peaked, the secondary and primary trimmers are tuned for maximum. The generator lead is then transferred to the grid of the previous tube. Generally, where the receiver has two i-f transformers, they will both be triple-tuned. The middle trimmer capacitor (tertiary winding trimmer) of the triple-tuned transformer is closed so that it is moderately tight (do not force the adjustment screw). The secondary and then the primary trimmers are adjusted for maximum output. The generator is transferred to the grid of the mixer tube. The middle trimmer capacitor (tertiary winding trimmer) of the first i-f triple-tuned transformer is opened (do not dislodge adjustment screw from nut). The secondary and then primary trimmers are adjusted for maximum output. The signal-generator lead is transferred to the antenna terminal of the receiver and the receiver tuning control is set at the low end of the dial to permit the r-f signal circuits to pass the i-f signal from the generator. (It may be necessary at this point to increase the output of the generator.) The second triple-tuned transformer tertiary trimmer is opened until maximum output is obtained. The first triple-tuned transformer trimmer is closed until maximum output is obtained. Neither the secondary or primary trimmers of the transformers should be readjusted.

A receiver incorporating triple-tuned i-f transformers may be aligned visually as follows: connect the f-m modulated signal generator high side to the grid of the last i-f tube and the low side to receiver chassis. Connect the oscilloscope terminals from the audio output tube plate to ground or across the detector diode-load resistor. The envelope of the voltage fed into the oscilloscope is built up across the detector load and it varies at an audio rate, namely, at the rate of the frequency modulation of the signal generator. To keep the pattern on the oscilloscope screen stationary, the oscilloscope horizontal deflection voltage (sweep voltage) must be synchronized with this audio rate. A synchronizing voltage, at the rate of frequency modulation of the signal generator must be fed from the signal generator to the horizontal input of the oscilloscope.

The signal generator is set to 450 kilocycles and a sweep width of plus and minus 15 kilocycles is used. The middle trimmer (tertiary winding trimmer) on the last triple-tuned transformer is tightened. The secondary coil trimmer is adjusted so that the top of the selectivity curve on the oscilloscope is centered on the resonance curve. The primary trimmer is then adjusted for maximum amplitude of the selectivity curve at the resonant point (center of selectivity curve). The output of the generator is reduced and the middle (tertiary) trimmer is adjusted for maximum amplitude and symmetry of the selectivity curve about the resonant center. The primary trimmer is then readjusted for maximum symmetry and amplitude. The signal generator is then connected to the grid of the mixer tube. The middle (tertiary) trimmer is unscrewed to a minimum capacitance position. (The output of the generator may have to be increased to obtain a large enough curve on the oscilloscope screen). Adjust the secondary of the first triple-tuned i-f transformer for maximum symmetry and amplitude. Adjust the primary trimmer for maximum amplitude. Reduce the generator output and adjust the middle (tertiary) trimmer for maximum symmetry and amplitude. The adjustments of the second triple-tuned transformer tertiary trimmer, the primary trimmer of this transformer, and the tertiary trimmer of the first triple-tuned transformer may be repeated carefully for more accurate alignment.

Resistance Loaded I-F Transformers

Receivers making use of resistance-loaded coils in the i-f transformers are really low-selectivity receivers rather than high-fidelity receivers. One of the characteristics of high-fidelity circuits is the non-acceptance of frequencies outside the bandpass. A high-fidelity

bandpass is thus distinguished not only by wide frequency acceptance and equal response for accepted frequencies, but also by sharp cutoff for frequencies outside the bandpass limits. A sharply tuned resonant circuit which has a resistor shunted across it, has its Q lowered by the loading effect of the resistor which decreases the selectivity of the circuit and widens its frequency response but does not set up sharp cutoff at the limits. Fig. 4 illustrates the response curve of a sharply tuned i-f transformer and Fig. 5 shows what occurs when this same transformer has one or both of its coils shunted by a resistor. Since the transformer is still single peaked at the intermediate frequency, alignment of a receiver incorporating resistance-loaded transformers is the same as for a sharply-tuned transformer except that trimmer adjustment will not be as critical.

Receivers making use of overcoupled transformers to obtain high-fidelity bandpass, where the overcoupling is the result of physical placement of primary and secondary coils with respect to each other, may be simply aligned. It is characteristic of fixed overcoupled transformers that the primary and secondary windings react on each other, so that the conventional method of aligning the two windings for maximum output at the i-f peak cannot be used.

One method of overcoming the reaction between the two windings is to shunt a resistor across one of the windings, while the *other* is being aligned, the value of the resistor depending on the degree of coupling, with a value of about 20,000 ohms being most commonly used for this purpose. The principle of operation is that the shunt resistor damps that tuned circuit of the transformer across which it is shunted so that it will not react on the other.
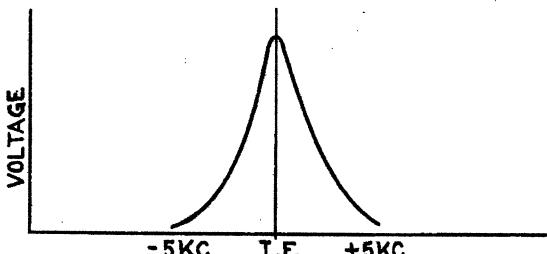
FIG. 4.—Response curve of sharply tuned i-f transformer.

After the one winding has been adjusted, the resistor should be removed, and it will then be possible to peak the other winding for maxi-
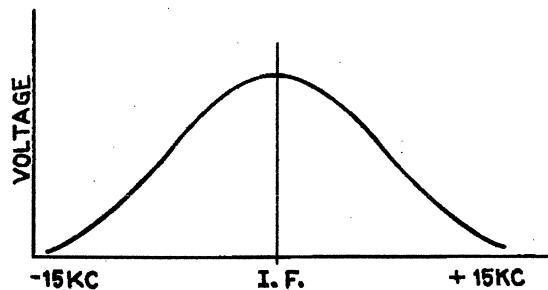
FIG. 5.—Response curve of resistance-loaded i-f transformer.

mum output. The removal of the resistor restores the response of the circuit to its normal characteristic. This method is repeated for any other resistance—loaded transformers in the i-f amplifier.

Aligning Detuned I-F Transformers

Receivers obtaining their high-fidelity characteristic by means of detuning the primary and secondary coils of the i-f transformers from the intermediate frequency may be aligned in either of two ways. In the first method, the secondary coil of the transformer is tuned to one of the double-hump peaks forming the flat-top frequency limits, and the primary coil is tuned to the other peak. To check the adjustment, the signal-generator frequency should be varied between the flat-top frequency limits, and the output meter reading noted for uniform output over the frequency limits.

The second method of alignment is to adjust first the secondary and primary of the transformer to the i-f peak then slightly detune each (detuning one above and one below) until the variation of the output-meter readings as the generator frequency is varied over the desired flat-top range, is very slight. The change in reading of the output meter should be the same on either side of the center intermediate frequency.

Variable Selectivity I-F Transformers

Receivers which use i-f amplifiers having variable selectivity (by means of a mechanical system of variable coupling) must be aligned with the selectivity (or fidelity) control in the

minimum-coupling or maximum-selectivity position. In this way, the interaction between the primary and secondary windings, which ordinarily makes a special procedure necessary, is avoided. After the i-f alignment is completed with the selectivity control in the sharp-selectivity position, the overall alignment should be checked with the control in the broad-selectivity position. The variation in the output-meter indication should be symmetrical, and the two peaks, which are obtained as the signal-generator frequency is varied about 10 kc either side of the i-f peak, should have the same height.

Receivers making use of stagger-tuned i-f transformers may be aligned by the following method. Where the receiver has two i-f transformers, one should be peaked for maximum output at about 3 kc above the i-f peak, and one for maximum output at about 3 kc below the i-f peak. The overall response should now be checked by slowly varying the frequency of the signal generator over a band of frequencies of about 10 kc each side of the rated peak frequency of the i-f amplifier. Where the receiver uses three i-f transformers, the first may be adjusted at the i-f peak while the second and third transformers may be detuned 3 kc above and below the i-f peak respectively.

It is possible to improve the tone quality of older-type receivers having sharply tuned i-f stages by realigning the i-f transformers to provide a flat-topped resonance curve. This is accomplished by first peaking each i-f transformer to exact resonance, then slightly readjusting each i-f trimmer so that the resonance peaks will fall alternately on each side of the exact intermediate frequency. This process causes some loss in gain, but this is usually unimportant, compared with the improved tone quality obtained. Use of an f-m signal and cathode-ray oscilloscope in this readjustment will give best results.

It should always be remembered when it is impossible to drive a signal through the receiver, and the receiver voltages are correct, that the receiver may be very badly out of alignment and thus not permit a signal to pass through. Where this condition exists, the signal generator should be connected to the grid of the last i-f tube and this stage aligned. It will then be possible to drive a signal through from the preceding stage, assuming, of course, that the

stage is not inoperative for some reason other than incorrect alignment.

The Low-Frequency Oscillator Adjustment "Rocking"

Just as the high-frequency oscillator trimmer determines the performance of the receiver over the high-frequency portion of the band, so the low-frequency oscillator trimmer determines the performance over the low-frequency end of the band. The method of making this adjustment is different from the usual manner in which trimmers are peaked for maximum output, in that a procedure commonly designated as "rocking" must be used.

This rocking adjustment is carried out in the following manner: The signal generator and receiver are tuned to the point near the low-frequency end of the band which is specified in the alignment data; to make this discussion more definite and easier to follow, we shall assume that the adjustment is being carried out for the broadcast band, in which case the signal generator would be set at 600-kc.

The receiver should be tuned for maximum output, and in general, the dial reading will not be exactly 600-kc, but may be off by as much as 10 or more kilocycles on either side. Whatever the dial reading—even if it is exactly 600-kc—the next step is to change the setting of the low-frequency oscillator adjustment slightly and then to tune the receiver for maximum output. If this procedure increases the output, the setting of the oscillator trimmer should be changed a small additional amount in the same direction, and the receiver again tuned for maximum output. On the other hand, if the movement of the oscillator trimmer in this same direction and the readjustment of the tuning control reduces the output, then a slight variation of the trimmer in the reverse direction should be tried, and the receiver tuning control should be readjusted for maximum output.

This procedure of alternately adjusting the oscillator trimmer and the tuning control should be continued until no further increase in output can be obtained, that is, until the displacement of the oscillator trimmer in both the clockwise and counter-clockwise directions and the accompanying rotation of the tuning control for greatest output is accompanied by a reduction in the output. The object of this

procedure is to arrive at that adjustment wherein the r-f circuits are tuned to the signal, and the oscillator frequency is higher than the signal frequency by the amount of the i-f peak, so that the greatest receiver sensitivity is obtained. It should be noted that, in general, the dial calibration will not be exactly correct, this is nevertheless the best possible adjustment.

Upon completion of this rocking adjustment near the low-frequency end of the band, the adjustments near the high-frequency end should be repeated. Unless the alignment adjustments were initially very far off, it will not be necessary to repeat the adjustments at the low-frequency end of the band.

Alignment Coupling Methods

One of the important considerations in the alignment of receivers is the method of introducing the alignment signal from the signal generator into the receiver. The method or methods used assume great importance for the following reasons:

Improper connection of the signal generator to the receiver may load a circuit, this often occurs when the generator is connected to a tube grid, which results in the detuning of the resonant circuit, particularly where the output capacitance of the generator is relatively high. Improper connections require the feeding into the receiver of an output above the capability of the generator; for instance, when the generator is connected for i-f alignment to the signal grid of a mixer tube whose detector coil has low impedance for the i-f frequency and so shunts most of the alignment signal to ground. Failure to connect a capacitor in series with the "hot" lead from the signal generator to the receiver (where no series capacitor exists in the generator output circuit) to block d.c. may result in damage to the generator, or may result in a short-circuited d-c grid path (if the generator is connected to a tube grid) since the signal generator output circuit may offer low resistance compared to the generally high-resistance grid circuit of tubes.

Removing the grid lead from the grid of i-f or r-f tubes when connecting the signal generator to the signal grid may affect the voltage distribution in the receiver and may result in the signal not having the proper bias, this is particularly important when receiver design is such that the grid is returned to a point in the receiver other than ground.

In cases where minimum bias voltage for the r-f and i-f tubes is fed through the grid rather than the cathode circuit, removing the grid lead and returning the grid to ground through the signal generator, would leave the tube with zero bias and cause an excessive and undesirable increase in plate and screen current. Failure to insert a series blocking capacitor in the negative leg of the signal-generator connection to the receiver may result in short circuits in certain types of ac-dc receivers. A shielded lead should be used to connect the generator to the appropriate point in the receiver, and this lead must be kept as far as possible from the grid leads of adjacent tubes or stray coupling may result. Failure to use the standard dummy antenna (I.R.E.) may result in undesired decrease in signal input to the receiver, stray coupling, loading effects, and pickup from external fields, and it may also be impossible to align the antenna trimmers properly on some receivers or the loop trimmers on other receivers, unless a dummy antenna is used to simulate normal input loading.

It must be recognized that where standard methods of connecting the generator to the receiver are not used, the possibility exists of leaving wires unsoldered or having cold-soldered joints (where wires are disconnected) and that much time will be required (particularly when aligning various makes of receivers) to analyze and determine connection points and devices and to make the proper connections.

It can be seen that a simple, standard method for connecting the signal generator to the receiver for alignment would be most desirable. Two methods of coupling the signal generator to the receiver which are simple, and that have already been standardized to a great extent are given below.

One standard method for feeding the generator signal into the receiver calls for connecting the signal generator high side to the receiver input terminals (after disconnecting the receiver antenna) through the standard I.R.E. dummy antenna, using a shielded cable. This is the only generator connection made and it is not disturbed during the entire alignment process. Since the dummy antenna serves as the coupling link between the generator and receiver, eliminating the necessity for careful

determination of coupling points and devices, it will be described in detail.

The dummy antenna consists of inductance, capacitance, and resistance of such values as normally occur when a receiver is connected to an antenna, and thus they simulate the characteristics of a receiving antenna. A standard dummy antenna has been adopted by the Institute of Radio Engineers for this purpose and the values chosen are capacitors of 200 and 400 mmf, an inductor of 20 microhenrys, and a resistor of 400 ohms, connected

Fig. 6.—I.R.E. standard dummy antenna.

as shown in Fig. 6. The output impedance of signal generators used for service work generally is unknown, and it often varies greatly with the setting of the attenuator. Little is gained then by using the standard dummy antenna, unless it is used together with a precision signal generator of known output-impedance characteristics. (The output impedance of the signal generator together with the impedance of the dummy antenna constitute the loading on the receiver input.)

For general use with the average service signal generator, the unit shown in Fig. 7 can be used as the dummy antenna. It will operate satisfactorily over the broadcast and some of the lower-frequency short-wave bands.
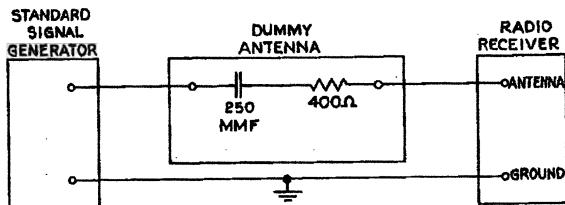
Fig. 7.—Alternative type of dummy antenna.

The dummy antenna should be constructed as a small independent unit to prevent coupling to other points in the receiver. The resistors and capacitors used must be non-inductive. The

stray capacitances between any two points must be so small as to be negligible at operating frequencies.

The effective values of the components should be within 10 per cent of the nominal values. The values of resistance, inductance and capacitance given above are the total values. The output impedance of the signal-generator attenuator circuit should be very small compared with that of the dummy antenna or should be subtracted from the values of the dummy antenna. The leads used in connecting the signal generator through the dummy antenna to the receiver, should be short so as to introduce very little voltage drop, and should be shielded to reduce external fields.

The standard dummy antenna functions in the following manner: At broadcast frequencies (540 kc to 1600 kc) the unit acts like an ordinary open-wire aerial which is resonant at approximately 2500 kc and with a capacitance about 200 mmf. At frequencies above the broadcast band, the impedance of the dummy antenna becomes constant at about 400-ohms. Under this condition it simulates a transmission line which is non-resonant. This standard dummy antenna therefore serves well for both broadcast and short-wave band alignment. The alternative dummy antenna illustrated in Fig. 7 serves well for broadcast-band alignment but fails for use on higher-frequency bands.

Fig. 8.—Box-type loop antenna.

Another method of coupling the signal generator to the receiver is recommended for use with receivers having loop antenna. Loop antennas used in home receivers are generally of two types. The "box" type loop shown in Fig. 8, consists of a number of turns of insulated

wire from 1 to about 20, wound in the form of a rectangular open box and supported on some sort of non-electrical framework (wood or cardboard) to hold the turns of wire in a vertical plane. The "pancake" or "flat" type loop shown in Fig. 9, is wound in the form of a flat
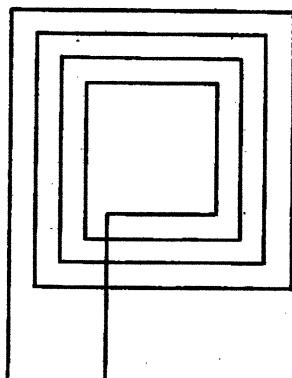
Fig. 9.—Pancake-type loop antenna.

spiral coil. The loop antenna functions as a result of the inductive action on the loop wires of the electromagnetic fields radiated from broadcast stations. A ground connection for the loop often is not required. Loops are used in home receivers merely as convenient antennas, since the main characteristics of the loop antenna, its directivity characteristics are of no interest or use (generally) in the home receiver.

On sets using loop antennas, coupling is best accomplished by feeding the signal-generator output into a loop of similar size and number of turns as the receiver loop and placing this loop near the one in the receiver. Such a transmitting loop is made easily by winding a few turns of No. 20 or 30 wire on four nails spaced in the shape of a square on a piece of plywood, and will serve more or less adequately for aligning most loop receivers.

No change need ordinarily be made when aligning the higher-frequency bands than increasing the distance between signal generator and receiver if necessary to attenuate the signal, since in many multi-band receivers, additional coils or loops are shunted across the main tuning loop to lower the inductance. For the long-range band, a loading coil is often placed in series, by switching, with the loop to build up the inductance to the required value for the range.

An increasing number of manufacturers are beginning to specify the use of the Hazeltine Model 1150 loop for alignment of their receivers. Fig. 10 illustrates the makeup of the loop which can be constructed without great difficulty. Its advantage lies in the fact that it is a permanent standard alignment tool which has been tested thoroughly and found to serve very satisfactorily as a signal-generator transmitting loop for alignment of all types of home loop receivers.

The Hazeltine Model 1150 loop consists of 3 turns of insulated No. 20 solid tinned copper
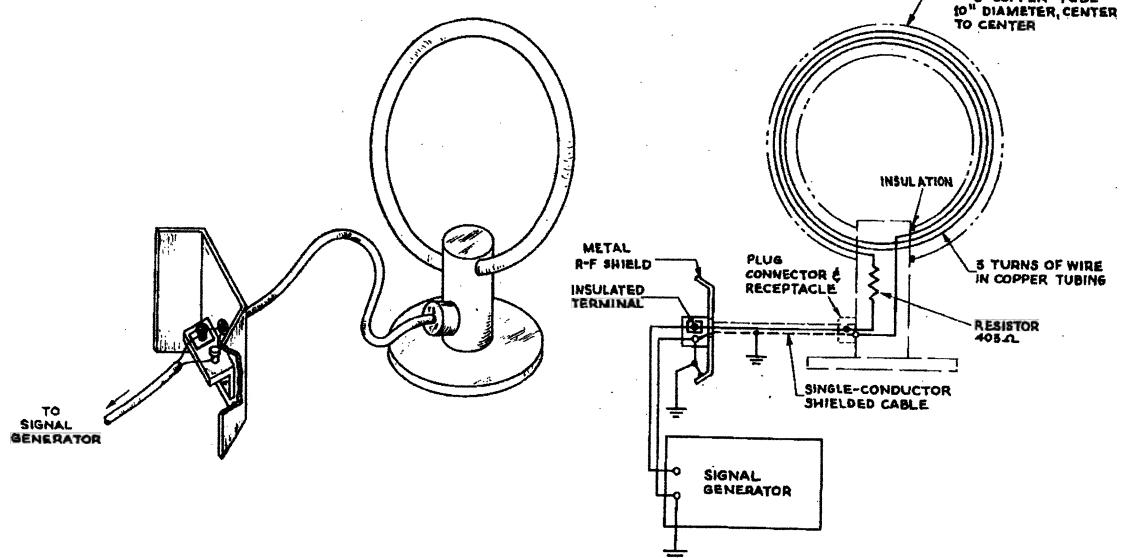
Fig. 10.—Hazeltine Model 1150 loop specified for alignment by various manufacturers. At the left is a drawing of the loop, and at the right is a schematic representation.

wire in a circular ⅜-inch copper tube that is bent to form a circle 10 inches in diameter. One end of the copper tube is insulated and the other end is connected to ground so as to prevent it from acting as a short-circuited turn, which would, of course, absorb energy. The connections of the loop are brought through the gap in the copper tube into a small housing at the base of the loop. The housing contains a 403-ohm resistor connected in series between the ungrounded end of the loop and the high side of the shielded cable going to the signal generator. A ¼-inch diameter, shielded microphone cable, 4 feet long, is connected at the loop by a microphone plug and jack. The other end of the cable is connected to "high" and ground of the generator through terminal posts on an aluminum shield large enough to prevent capacitive coupling between the receiver loop and the generator output terminals. The terminal on the shield to which the high side of the generator and loop are connected is insulated from the shield.

This transmitting loop creates an induction field of known strength into which the receiver loop is placed. When the receiving loop is situated 24 inches from the transmitting loop, then a field strength equal to one-tenth the generator output reading exists at the receiving loop. Use of a standard loop of known characteristics is useful when making stage-gain measurements, particularly when the antenna circuits include matching transformers, etc. For receivers having loops greater than 12 inches, the transmitting loop should be situated at a distance at least twice the maximum dimension of the receiving loop. The transmitting-loop characteristics are such that uniform current is developed through the loop up to a frequency of 20 mc.

Neither the signal-generator or receiver loops should be disturbed during alignment since this will break the effect of the continued decrease in the necessary signal-generator output as the circuits are lined up. Rotating loops should be fixed temporarily in some position during alignment so as not to be moved accidently.

THE DOUBLE SUPERHETERODYNE

A number of receivers have been manufactured in which the superheterodyne principle is employed twice in succession. Receivers, which employ this principle in a single unit, are known as double superheterodynes, of which the Stewart Warner Models 105 series are an example. In addition, any super-heterodyne such as one originally intended for use on the broadcast band, which is used in conjunction with a short-wave converter for short-wave reception, is another example of a double superheterodyne or a double-conversion system. The primary difference between these two types of double-conversion systems is that the former is a single unit embracing all of the elements, whereas the latter consists of two distinct units, namely, the converter as one and the receiver as the other.

The block diagram in Fig. 1 is an example of the general double-conversion receiver. To illustrate the operation of the receiver, the path of an 18-mc signal is traced through the receiver. At the same time, this will illustrate the frequency at which the circuits should be aligned. Of course, it is possible that other alignment frequencies may be specified, but the values given will serve as suitable examples.

The received 18-mc signal is impressed on the r-f amplifier and after being amplified, is fed to the short-wave detector or high-frequency mixer. In some cases, the r-f amplifier is not present, so that the received high-frequency signal is fed directly into the high-frequency mixer. The high-frequency oscillator generates a 19-mc signal, assuming that the first intermediate frequency produced is 1000 kc. As a result of the operation of the mixer

tube, the difference frequency of 1000 kc is produced and fed into the high-frequency intermediate amplifier; this 1000-kc signal constitutes the high intermediate-frequency signal. This completes the first frequency conversion.

After due amplification, the high intermediate-frequency signal is fed into the second mixer or broadcast detector. The design of this receiver is assumed to be such that the low intermediate frequency in the broadcast section is 460 kc. In order that this signal be developed, the low-frequency oscillator generates a voltage which is 460 kc higher than the signal fed into the broadcast mixer. This means that the oscillator frequency is 1460 kc. As a result of the operation of the broadcast mixer, the 460-kc intermediate-frequency signal is developed and fed into the low intermediate-frequency amplifier and, in the usual manner, reaches the second-detector tube, wherein the signal is demodulated and the audio voltage is fed into the audio amplifier.

It is significant to note that in a receiver of this type, only the r-f high-frequency mixer and high-frequency oscillator are variable tuned. The high-frequency intermediate amplifier, the broadcast mixer, and the low-frequency oscillator are fixed-tuned to frequencies which are independent of the signal being received. This becomes evident upon analysis of the circuit, since the frequency relation between the received signal and the high-frequency oscillator is always such that the difference frequency of 1000 kc is developed. Since the input to the high-frequency intermediate amplifier is a fixed frequency, the input to the broadcast mixer is likewise a fixed frequency irrespective
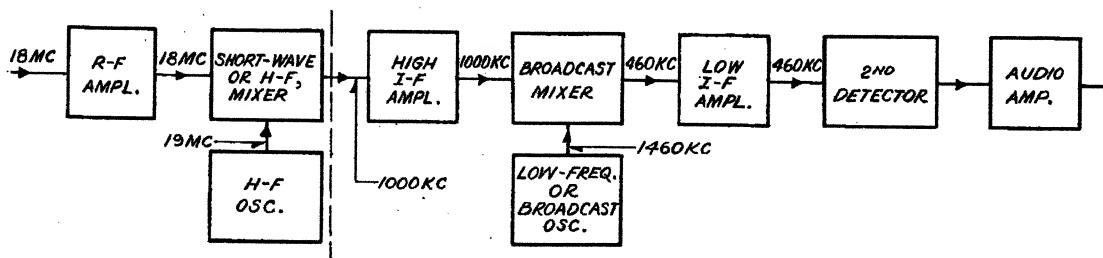
Fig. 1.—A typical double-superheterodyne receiver in which the two oscillators supply currents that produce intermediate frequencies of 1000 kc and 460 kc.

of the frequency of the original signal received; the low-frequency oscillator also can be a fixed frequency.

Alignment

As far as alignment is concerned, this double-conversion system can be divided into two separate sections and handled as if they were two separate superheterodyne receivers connected in tandem or series. For example, as can be seen in Fig. 1, that part of the circuit to the right of the dotted line constitutes a complete superheterodyne receiver with its input circuit tuned to 1000 kc and its i-f amplifier tuned to 460 kc. It is possible to view this system as being an r-f amplifier fixed tuned to 1000 kc and an oscillator fixed tuned to 1460 kc. Alignment would then be carried out at 1000 kc in the broadcast mixer, and alignment of the low-frequency oscillator would take place at 1460 kc. In turn, alignment of the low-frequency intermediate amplifier would take place at 460 kc. The order of alignment would be in conformity with what has already been said in connection with the conventional superheterodyne receiver.

Let us now consider this system to the left of the dotted line. In accordance with the frequencies mentioned, the high-frequency r-f amplifier, the high-frequency or short-wave mixer and the high-frequency oscillator to the left of the dotted line would be aligned over the normal tuning band—bearing in mind that the difference frequency out of the high-frequency mixer is 1000 kc. The alignment procedure here is similar to that described for the ordinary superheterodyne.

The circuit to the left of the dotted line in Fig. 1 is the equivalent of a normal converter system, when it is contained in a separate unit distinct from the rest of the circuit and it is

Fig. 2.—A double superheterodyne as used for an amateur 6-meter receiver.

aligned in a conventional manner. If, for example, this converter were used with a tuned-radio-frequency receiver, then the complete combination is an ordinary superheterodyne receiver, wherein the converter contains the ordinary mixer and oscillator, and the trf receiver supplies the intermediate-frequency amplifier and second detector.

In some instances, such converter systems include a wave trap in series with the antenna circuit, or at least in the antenna circuit. This wave trap is tuned to the output frequency of the converter, which in Fig. 1 would be 1000 kc. The reason for the trap is to eliminate interference which may arise because the 1000-kc amplifier circuit is within the broadcast band. In some instances, such interference is picked up by the lead which connects the converter detector to the high-frequency intermediate amplifier.

Why the Double Superheterodyne is Used

The reasons for using the double-superheterodyne process will now be discussed. As a practical example, let us suppose that it is desired to cover some high-frequency range, for instance the 6-meter amateur band, from 50 mc to 54 mc. If the intermediate frequency for this receiver were very low, such as the fairly common value of 455 kc, trouble would be experienced with images, since the input circuits cannot be made sharp enough to eliminate an interfering signal 910 kc from the desired signal. (The image signal is twice the intermediate-frequency value away from the r-f input signal.) If better discrimination against images is desired, the image-frequency must be further removed from the signal frequency and the only way of doing this is by making the intermediate frequency high, such as 5 mc or some such high value.

The selectivity of a superheterodyne receiver is determined mainly by the selectivity of the i-f amplifier. The selectivity of an amplifier operating at 5 mc, or some other such high frequency, is not very great in terms of broadcast reception, since at high frequencies even a sharply tuned amplifier would have a relatively wide bandpass. It becomes desirable to make the i-f amplifier itself a superheterodyne receiver, by introducing a second oscillator and mixer and having a second intermediate frequency of about 455 kc. By this
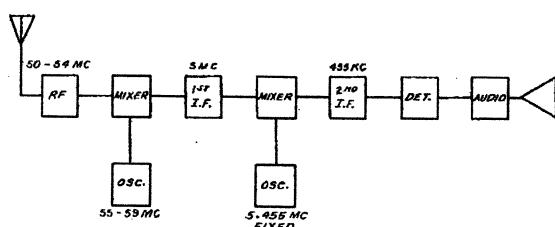
means, both the image-frequency discrimination and the desired selectivity may be secured. Fig. 2, illustrating a typical double-superheterodyne receiver, can be analyzed as follows, using a 6-meter amateur-band receiver as an example.

The desired signal is in the band of 50 mc to 54 mc, and the first oscillator is tunable from 55 mc to 59 mc, thus creating a constant difference frequency of 5 mc. This 5-mc signal passes through the first intermediate-frequency amplifier and is applied to a second mixer. Here it beats with the signal of the second oscillator which remains fixed at 5.455 mc and creates a second intermediate-frequency signal of 455 kc. This 455-kc signal passes through the second intermediate-frequency amplifier, is demodulated, amplified, and fed to the receiver speaker.

The advantages of the double superheterodyne include the possibility of having a high first intermediate frequency and yet retaining high selectivity, as mentioned previously. Further, additional gain can be obtained from the entire system, since the total gain is spread over amplifiers operating at different frequencies, so that the amplifiers do not react on each other so as to cause degenerative feedback or oscillation. The double superheterodyne, however, is subject to special difficulties which will now be described.

Difficulties Encountered

In the first place, while the single superheterodyne has the image-frequency phenomenon, the double superheterodyne has three images. The receiver, shown in Fig. 2, will be used as an example again. Since the first 5-mc i-f amplifier is followed by a second one having an intermediate frequency of 455 kc, it follows that besides a signal of 5 mc in the first i-f amplifier, there will also be a signal which is 455 kc above the second-oscillator frequency. In other words, assuming a 5.91-mc signal is present it will also pass through the second mixer and be converted to 455. Any signal reaching the first mixer, which in beating with the first oscillator can produce a signal at either 5 mc or 5.91 mc, may pass through the entire receiver unless the high (first) or low (second) i-f amplifiers are selective enough to eliminate them.

Assume that the receiver is tuned to a signal of 53 mc and that the first oscillator therefore is tuned to 58 mc. This will cause the desired signal to be converted to 5 mc and be passed on through the rest of the set. If a second signal is received at 63 mc and beats with the first oscillator, this signal will pass also through the receiver and would be called the first image frequency. Thus there are two possible signal frequencies which differ from the oscillator frequency by 5 mc. There are also two possible signal frequencies which differ from the oscillator frequency by 5.91 mc when the receiver is tuned to 53 mc and the oscillator is operating at 58 mc; these are —5.91 mc or 52.09 mc and +5.91 mc or 63.91 mc. As has been shown, the first and second oscillators finally will convert these four signals to 455 kc so that the three image signals (52.09 mc, 63 mc, and 63.91 mc) will interfere with the one desired signal at 53 mc. Fig. 3 illustrates the frequency spacing of the signals concerned and shows how the second and third images appear with respect to the first oscillator frequency.

To eliminate these undesired signals, the first r-f and mixer circuits must have sufficient selectivity to remove the undesired signals at 63.91 mc and 63 mc and the first i-f amplifier should have sufficient selectivity to remove the 5.91-mc signal which only could be caused now by an interfering signal at 52.09 mc. The receiver should be designed so that these conditions are fulfilled. Otherwise if a station happens to be transmitting at any one of the image frequencies, more than one signal will be heard at the same time or one station may appear at four points on the dial, if the tuning range covers all these images. A 63-mc signal may appear when the dial is tuned to 62.09 mc, 63.0 mc, 73 mc, or 73.91 mc if the receiver is not designed with sufficient selectivity.

Another trouble peculiar to the double-superheterodyne receiver is the appearance of
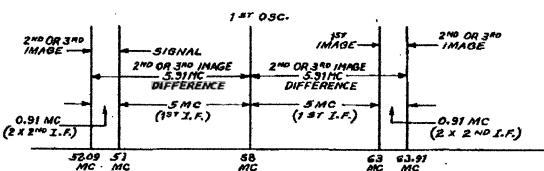
Fig. 3.—Frequency spacing of the signals that show how the second and third images appear with respect to the first oscillator frequency.

"ghost" signals, which are caused by the interaction of the two heterodyning oscillators. Although it is true usually that the oscillator coils are shielded and the various power-supply leads are filtered, the oscillator signal may travel from its proper path. The signal from the second oscillator, for instance, should go into the second mixer and should then not travel anywhere else. However, it is liable to appear at the input of the receiver or first mixer. This can happen through radiation, or by capacitive or inductive coupling, or through the common power-supply leads. Similarly, the signal of the first oscillator may appear at the grid of the second mixer. In addition, both of these oscillators may produce harmonics of the fundamental frequency, and/or, harmonics of the oscillator; signal and beat fundamental frequencies may be generated within one or both of the mixers.

Whenever the beat between two harmonics of the first and of the second oscillators is equal to the signal frequency or one of its image frequencies, or is equal to the first intermediate frequency or the second intermediate frequency, this beat will pass through the receiver just like any other signal, and will be evident in the loudspeaker as an unmodulated carrier. If the intermediate frequencies have been improperly chosen, there may be very many of these ghost signals resulting from harmonics of beat frequencies, and beats between harmonics and beats, etc. This is one of the greatest problems in the design of the double-superheterodyne receiver.

Causes of "Ghosts"

In order to illustrate this phenomenon more fully, the double-superheterodyne receiver described above will be examined for the possible causes of ghosts. The signals of both oscillators may appear at any point in the receiver (unless there is extreme care in designing the receiver) so that the tenth harmonic of the second oscillator fundamental frequency (5.455 mc) which is 54.55 mc, may cause trouble. The eleventh and higher order harmonics are beyond the limit of the receiver tuning range and the receiver will just miss their interference, particularly since the higher the harmonic, the smaller is its intensity. At any rate, the tenth harmonic mentioned above as

an example is within the receiver tuning range and may beat with an r-f signal to produce an interfering signal at the first or second intermediate frequency. It is the aim of the designer, then, to choose the intermediate frequencies so as to minimize the number of ghosts.

Now let us consider the second harmonic of the first oscillator. When the receiver is tuned to 52.505 mc, the first oscillator is operating at 57.505 mc and the second harmonic of the first oscillator is 115.010 mc. This will beat with the 21st harmonic of the second oscillator (114.555 mc) to cause a 455-kc beat which passes through the second i-f amplifier, and therefore will be heard as a ghost signal.

Since the higher harmonics of the oscillators are considerably weaker than the fundamental and second-harmonic frequencies, it is especially important to avoid a design in which the lower harmonics of the two oscillators can form an interfering beat, for this would then have great interfering strength. In the above example this did not happen, for this particular design has only a few possible ghosts. However, if the first intermediate frequency had been much lower and the tuning range greater, there could be very many ghosts.

In another receiver, suppose that in order to cover high frequencies from 3500 kc up, the first i.f. was chosen at 1600 kc and the second at 455 kc. Then the second oscillator would operate at 2055 kc and its second harmonic, 4110 kc, would appear as a ghost signal when the receiver is tuned to 4110 kc. When the receiver is tuned to 5020 kc, the first oscillator is operating at 6620 kc and will beat the fourth harmonic of the second oscillator (8220 kc) to form a 1600-kc signal, which passes through the receiver and appears in the loudspeaker as a ghost signal. There may also be many other ghosts in this receiver. It should now be clear that many of these ghosts may exist when the first intermediate frequency is low and the receiver tuning range is large.

The existence of ghost signals depends on oscillator harmonics, and the presence of these harmonics in places where they do not belong. This gives a clue to the methods for eliminating them. First, the oscillator circuits should be designed so as to have a minimum number of harmonics, which can be done by proportioning the oscillator tickler coil (where used) so

that the tube barely oscillates at all points in the tuning range. Second, shielding and filtering should be thorough, so as to prevent the oscillator signals from getting to the wrong places. The third remedy is to choose the intermediate frequencies in such manner as to make the ghosts fall outside of the tuning range.

The double superheterodyne is in practical use despite these problems. When a high-frequency converter is used ahead of an ordinary superheterodyne receiver, the combination becomes a double-superheterodyne receiver. When a multi-band superheterodyne receiver is used as an i-f amplifier, for instance at 8.25 mc, for the sound signals of a television broadcast,
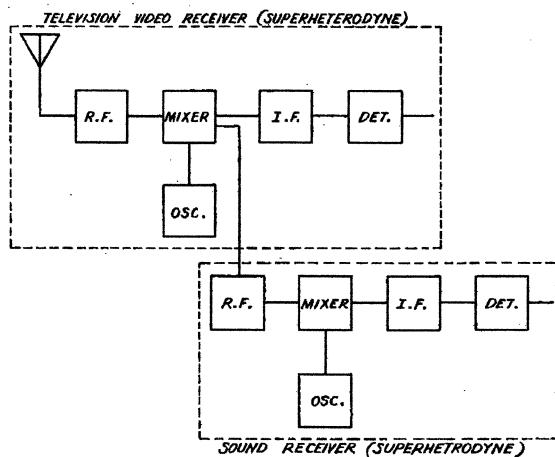
Fig. 4.—Block diagram for a double superheterodyne for video and sound reception.

the television receiver r-f circuits and the multiband superheterodyne receiver constitute a double superheterodyne. Fig. 4 is a block diagram of such a combination.

G. E. Translator, Type JFM 90

Sometimes, in order to eliminate ghost signals, receivers are designed in which the signal of a single oscillator is used twice, so that there will not be two oscillators which may beat with each other. One receiver using this system is the frequency-modulation tuner known as the General Electric Translator, Type JFM 90. This receiver covers the frequency-modulation band from 42 mc to 50 mc. Fig. 5 is a block diagram of the operation of this receiver. From the antenna, the input signal goes to the mixer grid, the mixer circuit being tunable from 42 mc to 50 mc. The oscillator is variable from

18.85 mc to 22.85 mc and heterodynes with the incoming signal.

This mixing action is different from conventional superheterodyne operation in that a fixed intermediate frequency is not produced by beating the r-f input signal against the receiver oscillator. The mixer-oscillator tracking is not constant, that is, when the mixer is tuned to 42 mc, the oscillator is tuned to 18.85 mc thus producing an i.f. of 23.15 mc, and when the mixer is tuned to 50 mc, the oscillator is tuned to 22.85 mc thus producing an i.f. of 27.15 mc.

The first i-f circuit is variable in tuning, and is controlled by a capacitor section on the same shaft with the oscillator and mixer tuning capacitors. These capacitors have been so proportioned that they are always tuned to the proper frequencies at all points of the dial.

In this receiver, the first i-f circuit has been cut to the barest essentials, for it consists of just one tuned circuit and does not contain an amplifier stage. An amplifier tube is not absolutely essential here, so that for purposes of economy it is omitted. The first mixer feeds directly into the second mixer through this single-tuned circuit which is tuned to the first intermediate-frequency band. Although a single-tuned circuit is not very selective, it does reduce the image-frequency signal (which would cause interference after the second mixing operation) as well as other undesired signals at frequencies far removed from the first intermediate-frequency band.

The second tube in the receiver is another mixer. The i-f signal, whose frequency is between 23.15 mc and 27.15 mc, beats in the second mixer with the original oscillator frequency (whose value is between 18.85 mc and 22.85 mc). This original oscillator frequency has been able to pass through the first i-f amplifier, which is not very selective and will pass the
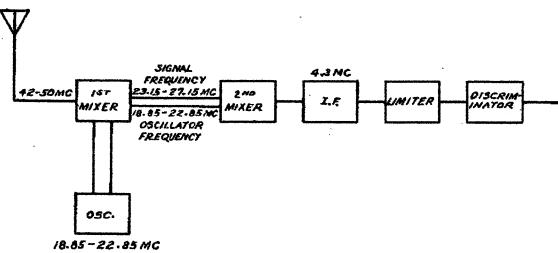
Fig. 5.—Block diagram of the General Electric Translator Type JFM90 used as an f-m tuner.

oscillator signal, since the oscillator frequency does not differ much from the first i-f signal. The frequency of the beat after the second mixing operation is 4.3 mc and is constant regardless of the position of the tuning dial. The second i-f amplifier is tuned to 4.3 mc, and the rest of the receiver follows customary procedure for f-m converters.

The reason for using a double superheterodyne at all might seem difficult to understand, since no amplifier is used in the first i-f circuit. Sufficient amplification must be achieved if full advantage is to be made of the possibilities of f-m reception. Usually, it is difficult to get sufficient amplification at high frequencies. In the case illustrated, the 4.3-mc second i-f circuit could not have any more gain without the danger of making it unstable. Of course, an r-f stage could give additional gain. However, it is impossible to get as much gain out of a tube used as a high-frequency r-f amplifier as it is from the same tube when it is used as a mixer since its plate circuit in the latter case is tuned to a lower frequency. An additional amplifier stage in the first i-f circuit would necessitate at least one more tuned circuit, which, as explained above, must be variable and its tuning capacitor must be ganged to the tuning capacitors of the other circuits. Accordingly, it is omitted in order to simplify the receiver design.

A Fixed-Frequency Oscillator

The principles described in this article can be used in many ways. One special receiver employing several of the features discussed above was described by Braulio Dueno in "Electronics" for August, 1939. A simplified version of the conversion section of the receiver is shown in Fig. 6, and its operation may be analyzed as follows: the r-f input signal from the antenna is applied to the grids of both 6A7 tubes. In mixer tube No. 1, the r-f signal is mixed with a local oscillator signal at a fixed frequency of 450 kc. In the plate circuit of this tube is a tuned circuit, whose capacitor is ganged with the mixer tuned-circuit capacitor, which selects from the various frequencies appearing in the plate circuit, the difference frequency, which is equal to the signal frequency minus 450 kc. This difference frequency is applied to the oscillator grid of the second mixer

and is again mixed with the original input signal, so as to produce a beat equal to their frequency difference, which is always 450 kc. The intermediate-frequency amplifier is tuned to 450 kc, while the rest of the receiver follows standard design.
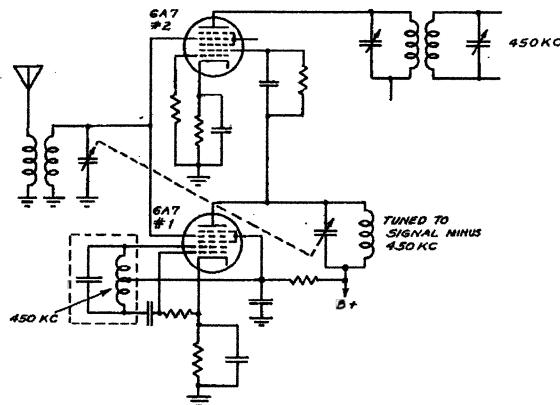
Fig. 6.—Simplified schematic of the conversion circuits of a double superheterodyne wherein the signal from the antenna is applied to the grids of both 6A7 tubes. The difference frequency appearing in the plate circuit of the No. 1 tube is applied to the oscillator grid of the No. 2 tube, where it is mixed with the antenna input signal, the output being applied to the i-f input transformer.

The advantages claimed for this type of superheterodyne are: no troubles due to inefficient operations of the oscillator at very-low and very-high frequencies, and no trouble due to frequency drift of the oscillator, since the constant-frequency oscillator can be designed to have great stability. There is the possibility, however, of the oscillator signal getting into the i-f amplifier directly, and thus good filtering and shielding of the oscillator and i-f amplifier are required. Further, any interfering signal reaching both mixers which has not been filtered out by the mixer tuned circuit and the first i-f circuit (plate circuit of first mixer), will cause a 450 kc i-f signal which cannot be removed by the second i-f amplifier. In other words, the selectivity of the receiver is determined by the input circuit and the variable i-f amplifer.

Double Superheterodyne with a Super-regenerative Second Detector

A special receiver for reception of short waves was described by Ross A. Hull in "QST" for November and December, 1935. Fig. 7 is

a block diagram of the conversion operations.
The receiver is a double superheterodyne with
a super-regenerative final detector, but in this
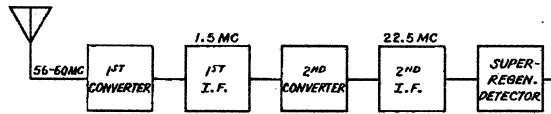
FIG. 7.—Block diagram of a double superheterodyne
with a super-regenerative second detector.

set the intermediate frequency used after the
first mixer is lower than the second intermediate
frequency. This is done to obtain first the ne-
cessary selectivity; then the frequency is
changed to 22.5 mc so that a super-regenerative

detector would function more efficiently. If
the super-regenerative detector alone is used,
the selectivity is insufficient, as the super-re-
generator is not an efficient detector at low
frequencies; therefore, it is necessary to change
the frequency back to the high-frequency region
in order to obtain both high sensitivity and
high selectivity.

An AM-FM Double Superheterodyne

Somewhat indicative of a possible trend to
increased use of double-superheterodyne action
in the future, are the production of several
double-superheterodyne receivers by Crosley
for 1947. The Crosley Models 86CR and 86CS
receivers have been designed ingeniously so
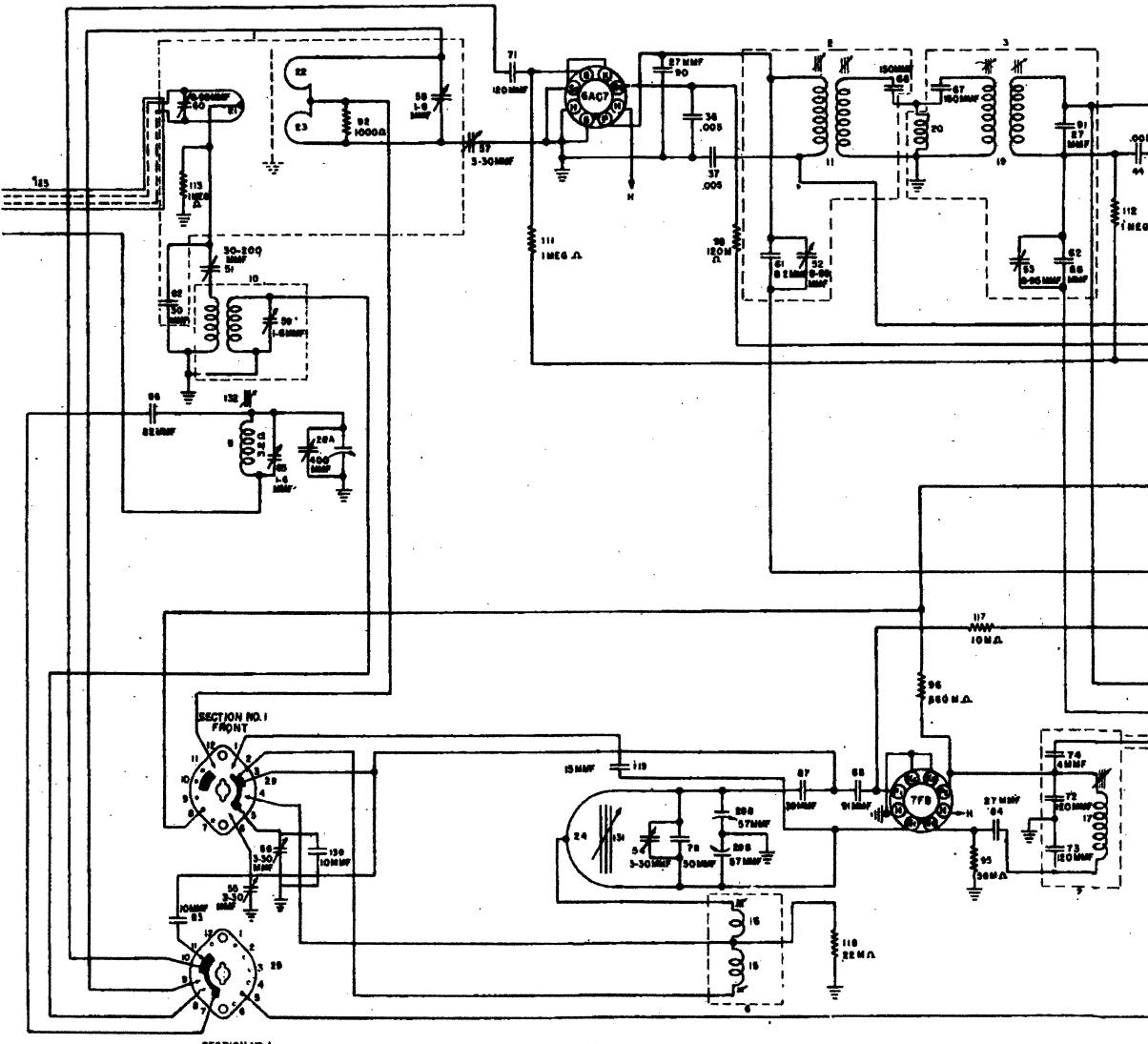
that a total of fifteen separate functions are performed by eight tubes. These models have double-superheterodyne action on the broadcast and short-wave a-m bands and incorporate a single superheterodyne f-m circuit for reception of the 88 mc to 108 mc f-m band.

Fig. 8, the schematic of this receiver, shows that this receiver generally follows the block outline of operations for a typical double-superheterodyne receiver, as illustrated in Fig. 2; however, there are some interesting differences. No r-f amplifier stage is included before the first mixer and there is also no i-f amplifier stage after the first mixer, so that the first mixer feeds through a double-tuned i-f transformer to the second mixer. The frequency of

5825 kc is used as the first intermediate frequency so that it is impossible to have trouble with image frequencies on either the broadcast or short-wave bands. This is so because the image frequency (twice the i-f value away from the signal frequency) is completely out of the range of both the broadcast and short-wave bands. Any decrease in gain or selectivity resulting from using the comparatively high first i-f of 5825 kc is more than compensated for by the use of the comparatively low second intermediate frequency of 167.5 kc. To guarantee that there will be no interference from signals at or near this frequency of the first i.f., an i-f wave trap adjustment (trimmer capacitor 85), resonant to 5825 kc, is incorporated in the a-m
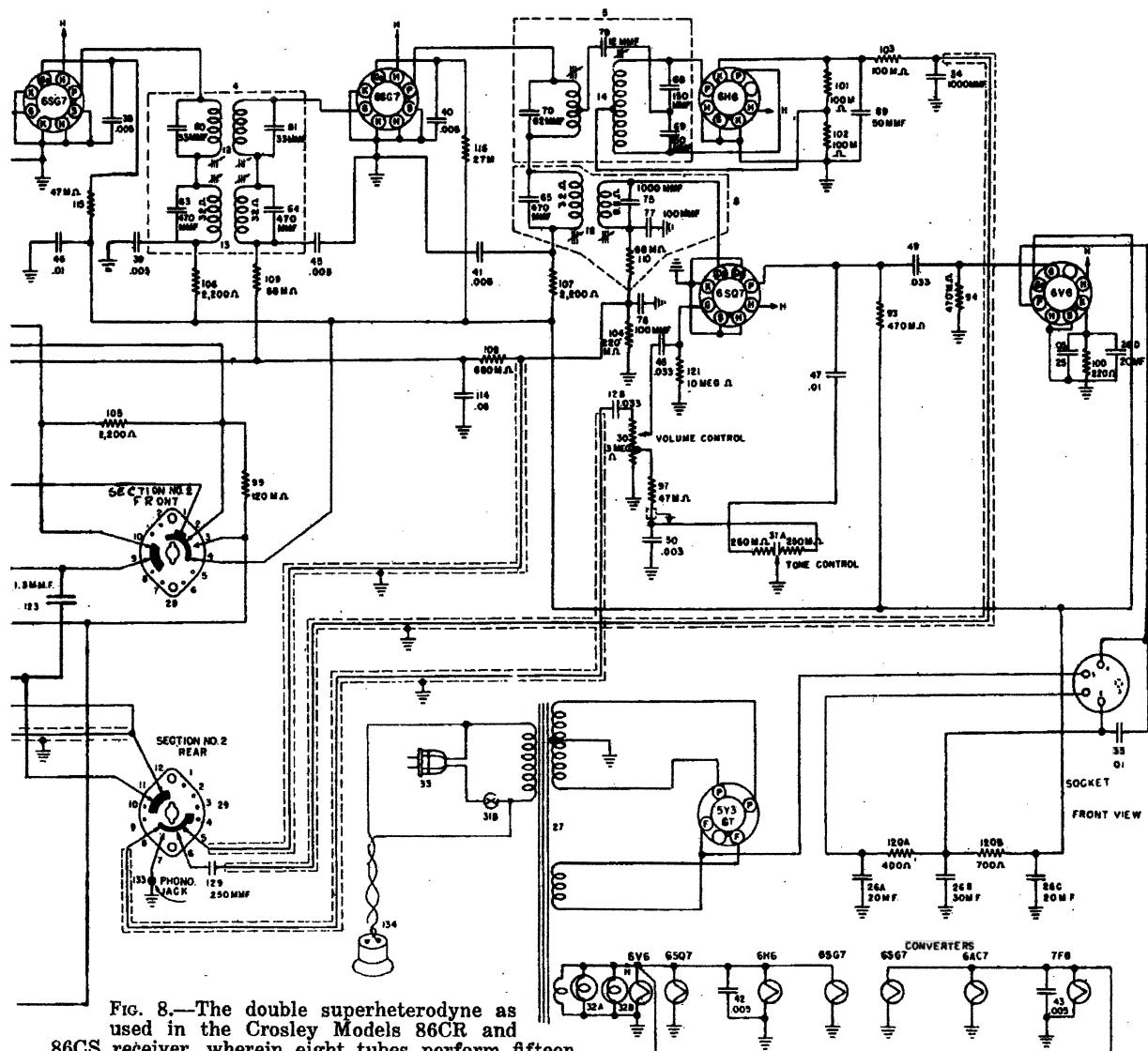
FIG. 8.—The double superheterodyne as used in the Crosley Models 86CR and 86CS receiver, wherein eight tubes perform fifteen functions.

antenna circuit. Note that one section of the 7F8 twin triode serves as the variable first oscillator, while the other section of the tube functions as the fixed second oscillator, pro-ducing an oscillator frequency of 5992.5 kc. This latter frequency beating with the first i.f. of 5825 kc produces the second i.f. of 167.5 kc at the output of the second mixer.

HOME RECORDING FEATURES

This section will discuss some of the different combination sets on the market which contain means for home *recording*. The important subjects that will be considered are the different types of recording—such as radio recording and microphone recording—methods of monitoring the volume of recording, and any unusual circuit features that deal solely with recording. In most of the recording combination sets that we have examined, a portion of the audio stage of the regular radio section of the set was used in recording positions through special switch selections. Since the complete schematics of many of these sets are extremely complicated (as the Packard Bell Model 1052), we have broken down the switching arrangements of such sets completely, thus making it much easier to trace each individual switch change. These breakdowns have also been made for the Wilcox-Gay Model 6B10 and the Spiegel Model TR-640. The recording sets discussed are analyzed from the viewpoint of understanding how the electrical components function with respect to the operation of the different recording positions. By electrical components we mean resistors, capacitors, inductances, transformers, tubes, and the like. Since each set is unique in the arrangement of circuits we will examine a few of the combination models individually.

RECORDIO MODELS 6B SERIES

In the Wilcox-Gay Recordio Models 6B10, 6B20, 6B30, and 6B32 there are two different

chassis, an early model and a late model, which are identified by their serial numbers. The schematic differences between the early and late models are very slight, so for the purpose of this discussion the later model only will be considered here. In this receiver are two different types of recording: radio recording and microphone recording. Each type is analyzed separately.

Record Radio

The "record radio" position of this model is shown in Fig. 1. This position is the more important one because it utilizes the *complete section* of the radio-broadcast position plus additional elements that change the electrical characteristics of the set to afford a means of successful recording. When the "record radio" push button is depressed the following switching arrangements in the set are of importance. First, the diode plate, pin 5 of the 6SQ7GT/G, second-detector-avc first-audio tube becomes operative. (In the regular radio-broadcast receiving position this diode plate does not function since no signal is applied to it.) Secondly, a high-valued resistance of 3.3 megohms *(R19)* is inserted between the output of the 6SQ7GT/G tube and the input to the second-audio tube which is one triode section of the duo-triode 6SN7GT tube. To complete this switching arrangement there is incorporated a recording amplifier, which is the other triode section of the 6SN7GT tube and it is used in conjunction with the cutting head. If the
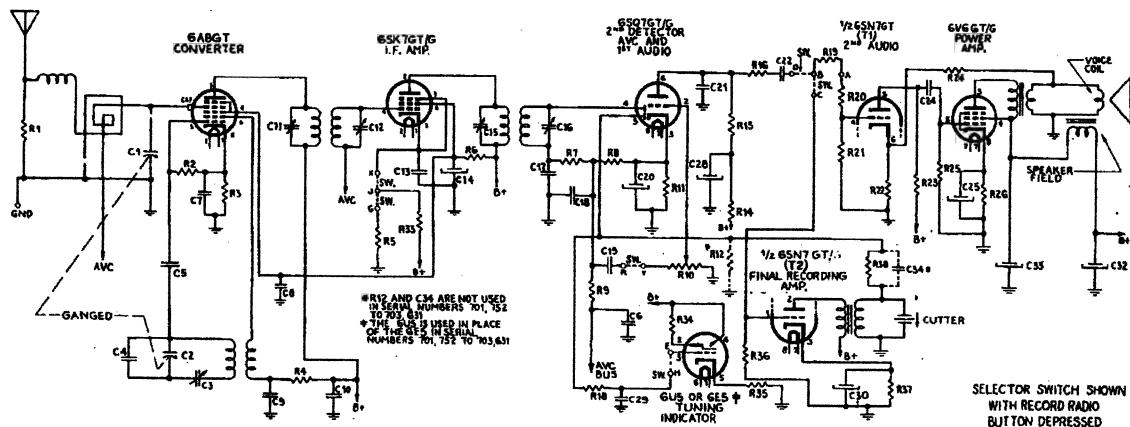
Fig. 1.—The complete section of the radio-broadcast portion of the Wilcox-Gay Recordio Models 6B series is used with a recording amplifier when the set is switched to the "Record Radio" position.

"radio-broadcast reception" *clarified schematic* (Wilcox-Gay, Page 15-2) is examined with respect to the "record radio" *clarified schematic* Fig. 1 in this article, the new components brought into play will be evident immediately.

Switching Features

The simple switching arrangement incorporated in this model has a number of special features that are clearly shown in the "record radio" position, Fig. 1. First, nothing is changed in the r-f section of the set. The only change in the avc-second detector section is that the avc signal is omitted from the indicator tube. The diode plate, pin 5, of the 6SQ7GT/G is connected through resistor *R18* to the grid of the indicator tube. Attached to this diode plate is a lead coming from the high side of the secondary of the recording output transformer through *R38*. Coming off the output audio section of the 6SQ7GT/G tube is a lead going to the grid of the 6SN7GT recording amplifier (T2), through *R16, C22,* and a switch. With the exception of this switching, no other changes are made to the circuit other than those already existing for regular radio broadcast reception.

An important feature in a well-designed recording set is the monitoring of the audio signal to be recorded. In other words, there should be some method of letting the operator of the set know when the correct audio signal voltage is impressed across the cutting head.

With all this in mind, refer to the complete *"clarified schematic"* diagram of the set in the "record-radio" position, Fig. 1, and trace a signal through the set. With the set tuned in to a particular station, the received signal goes through the proper heterodyning procedure and is converted into an i-f signal which goes to the diode plate, pin 4, of the 6SQ7GT/G tube. Here the signal is detected, and it emerges from the junction between resistors *R7* and *R8* as the modulated envelope of the r-f signal, which is the actual audio signal of the set. At the junction point mentioned, this audio signal has already gone through an r-f filtering network, *R7* in conjunction with *C17* and *C18*. From here the signal goes through an audio filtering network of *R9* and *C6* and emerges as the avc (d-c) signal which goes to

the control grid of the 6A8GT converter tube and the 6SK7GT/G i-f amplifier.

This audio signal is also impressed across the potentiometer, *R10*, which serves the purpose of a volume control, regulating the amount of audio signal fed onto the grid of the 6SQ7GT/G first-audio amplifier. This audio signal being amplified by the triode section of the tube appears at the plate (pin 6) of the 6SQ7GT/G tube. From here the signal is fed through *R16* and *C22* to point *D* on the push button switch. From point *D* the audio signal follows two paths: one is through pins *D* and *B* of the push button switch, through the 3.3-megohm resistor *(R19)* and 470,000-ohm resistor *(R20)* to the control grid of *T1*, the second-audio amplifier which is one triode section of the 6SN7GT tube; the other path is through the same switch from terminals *D* to *B* to *C* and then down to the control grid (pin 1) of *T2*, the final recording amplifier which is the other triode section of the 6SN7GT tube.

The most important factor in tracing the audio signal throughout this recording position is to know the exact paths the signal does take and why. The 3.3-megohm resistor *(R19)* is chosen to limit the amount of audio signal impressed onto the control grid of *T1*. The reason for this is that the sound coming out of the loudspeaker has to be muted so that the majority of the audio signal is used for recording and not for loudspeaker output. In this way the loudspeaker is used for monitoring the program to be recorded so that the operator will know what program he is recording.

The rest of the audio signal, as mentioned, is impressed across the grid of *T2*, the final recording amplifier. This latter amplifier increases the audio signal to a predetermined value (due to its amplification characteristics) and the output from the plate of the recording amplifier is impressed across the primary of the recording output transformer. By transformer action, the output audio signal from the primary is impressed across the secondary of the recording output transformer, and in turn this audio signal is further impressed onto the cutting head of the recording arm.

Monitoring the Audio Signal

The amount of audio voltage that is impressed onto the cutting head determines how

deeply the cutting head will cut. In order to know the proper amount of recording (i.e. to make sure that the cutting arm does not over-cut or undercut the record being made) an indicator tube is used. In the model being discussed, the tube used is either a 6E5 or 6U5. This set is so designed that when a station is tuned in properly and the volume control adjusted correctly, the eye of the indicator tube will just about close on *the peak passages* of the audio signal. A feature of this receiver is the way in which the indicator tube is made to react in such a manner. When the indicator tube is used in the regular radio-broadcast-receiving position the avc is applied to its grid and the tube then functions as an r-f tuning indicator for different stations. With the record-radio push button depressed, however, the regular avc is disconnected from the grid of the indicator tube and a new connection is made to the grid through points E and H of the push button switch. If the circuit is examined, it will be found that this grid is connected through resistor $R18$ to pin 5, the other diode plate of the 6SQ7GT/G tube, and also to the high side of the secondary of the recording output transformer through resistor $R38$. These new connections to the indicator tube afford a visual means of knowing when the optimum amount of volume is being recorded.

The operation of this section is as follows: Some of the audio output that appears across the secondary of the recording output transformer is taken off the high side of this transformer through resistor $R38$ (6.8 megohms). This signal (the strength of which is controlled by $R38$) is fed back to the diode plate, pin 5, of the 6SQ7GT/G tube. Rectification occurs and the pulsating output signal from this diode plate is filtered by the $R18$-$C29$ audio-filter combination and it is then applied as a d-c signal onto the grid of the indicator tube. This d-c signal is an automatic control signal that indicates the proper tuning of the set and also that the proper amount of audio volume is being impressed on the cutting head.

The regular avc signal keeps the level of the set constant by controlling the output of the r-f section. It then follows that the audio signal developed across the volume control is also constant — this assumes that the set has been tuned properly. The volume control, therefore, regulates the amount of signal fed into the

first-audio grid and hence controls the output from the first-audio tube. This output then splits up as stated before, a part of it going to the grid of the second-audio amplifier and part to the final recording amplifier. If the set is off tune, the audio voltage output from the 6SQ7GT G tube will be lower and hence the input to the recording amplifier will be less. This will then make the signal voltage fed back from the recording output transformer smaller, being shown on the indicator tube by the eye of the tube remaining open. For proper operation, the eye should just about close, and although the regular avc is disconnected from the indicator tube, this tube still serves as an r-f tuning indicator. By the proper choice of the feedback resistor $R38$ and the $R18$-$C29$ audio-filtering network, the input to the grid of the indicator tube will be constant when the eye is just about closed. *Therefore, when the volume control is adjusted to a point where the eye almost closes (or just closes on the peak passages of a signal), the set is adjusted properly for recording.* It should be remembered that the loudspeaker output in this model is not used as a recording level. The sound issuing from the loudspeaker indicates to the operator of the set what program he is recording.

Record Microphone

A *clarified schematic* for the "record microphone" position is shown in Fig. 2. When the "record microphone" push button is depressed, the *only new* section that is switched in is the microphone in conjunction with the 6J7GT microphone amplifier tube. The rest of the components that comprise this position are also used in the record-radio position of the set. Studying the circuit a little more carefully,


Fig. 2.—When the "Record Microphone" push button is depressed in the Wilcox-Gay Recordio Model 6B series, the microphone and the microphone amplifier is switched into the circuit as shown in Fig. 1.
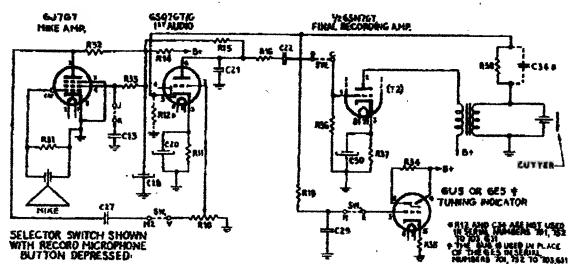
it will be found that the circuit accompanying the recording amplifier *T2* (one-half of the 6SN7GT tube) and the indicator tube are *exactly* the same as in the "record-radio" position but that the circuit of the 6SQ7GT/G tube has been changed. This change is that the diode plate (pin 4) circuit is made effectively inoperative. Since this section of the tube is only used for broadcast detection and avc action, it is then obvious that in the position under discussion, it is not needed because the r-f section is not in operation. The indicator tube in this position is used as a volume-level indicator for microphone recording the same way it was used in the "record-radio" position. In order to understand the way this set works on microphone recording it is best to trace the signal from the microphone to the recording mechanism.

The 10-megohm resistor, *R31,* across the microphone serves a dual purpose: it acts both as a load to the microphone and as a grid-leak resistance to the 6J7GT amplifier. The microphone used is a crystal type and as characterized by most of these types, the required load impedance is usually several megohms, depending on the particular microphone used. The resistance is made high in this case (10 megohms) because it is desired to have a good low-audio-frequency response and in order to achieve this, the load on the microphone should be as high as possible. After sound enters the microphone, it is converted into its electrical equivalents and impressed onto the control grid of the 6J7GT tube as an audio signal. This signal is amplified by the 6J7GT tube and it is then impressed across the volume control, *R10.* The variable arm of the volume control is connected to the grid of the triode section of the 6SQ7GT/G tube and the signal is further impressed onto this grid circuit through the volume control. This section acts as a second audio-voltage amplifier but the amount of signal impressed onto this latter grid is controlled and, therefore, its output is likewise controlled.

Tracing the Audio Signal

The audio signal that emerges from the plate of the 6SQ7GT/G tube follows *only one path* as compared with the two paths it took in the "record-radio" position. Here the total output is impressed onto the grid of the final re-

cording amplifier *T2* (one-half of the 6SN7GT tube) through pins *D* and *C* of the push button switch. The rest of the recording circuit is exactly the same as in the "record-radio" position. The indicator tube is similarly used as a *visual* means of informing the operator when the proper amount of audio signal is being impressed across the cutting head of the recording mechanism. In other words, when the eye of the indicator tube just about closes on the peak passages of the audio signal, then the volume control is set correctly. The important thing to bear in mind in microphone recording is that in using the microphone, one should talk (or sing or whatever the type of recording desired) into the microphone at a definite distance from it. In other words, the person who uses the microphone should not vary that distance because it will also vary the degree of volume cut into the record. It should be obvious to all that the reason the loudspeaker is not used for monitoring (as it is in the "record-radio" position) is because the *voice* of the individual using the mike is his own means of monitoring what goes onto the record. In the "record-radio" position, this is not so because one has to know what type of program is being received.

In the Wilcox-Gay Model 6B10, a tone control is used in the grid circuit of the second-audio amplifier *T1* (one-half of the 6SN7GT tube). (This can be seen by examining the *clarified schematic* for the "radio-broadcast reception" position in the Manual on Wilcox-Gay, page 15-2). In either recording position of this set, the tone control is switched out of the circuit so that there will be no possibility of having all bass recording if the tone control happens to be left in the bass position. Bass recording is an undesired feature because it really prevents the record being made from an input having the desired characteristics of high-fidelity recording.

PACKARD BELL PHONOCORD MODEL 1052

In the Packard Bell Phonocord Model 1052, there are three different types of recording, namely, radio recording, microphone recording, and mixed-program recording. The last type of recording is a combination of radio and microphone recording. This model is also a two-band set operating on the broadcast band

and the 6-18 mc. short-wave band. Both the "radio-record" and "mixed-program record" can be used on either band; however, for discussion purposes, reference will be made to only those recordings on the broadcast band.

In the previous discussion on recording, it was found that an indicator tube was used as a visual means of informing the operator when he had obtained the proper degree of cut onto the record. In the Packard Bell Phonocord Model 1052 there is *no* visual or manual method of controlling the degree of cut, *all* recording in this model being automatic. In other words, an automatic control circuit is incorporated which insures that the audio signal impressed across the cutting head of the recording mechanism is always regulated in volume. This automatic control is accomplished by a "volume-compression" circuit which is used on all recording positions of the set. Since this is true, it will be best to analyze a typical compression circuit before starting any separate discussion of the different recording positions of this set.

Volume Compression

In many different types of audio circuits, it is sometimes desired to control the volume range automatically because too high a gain and overloading effects may seriously damage some of the equipment used, and too low a gain will make for a poor response. This control is primarily accomplished by a volume compression circuit, which in brief, is one that reduces the audio gain when the signal that is to be amplified is large and increases the gain when the signal is weak. Volume compression is used in many different types of equipment and it finds one of its greatest uses in recording circuits. Before going into the actual compression circuit used in this model, it will be best to consider it from a generalized viewpoint.

A simplified volume compression circuit is shown in Fig. 3. Here tube *No. 1* serves as an audio amplifier and supplies an audio output signal from its plate circuit to two different tubes. Part of the signal goes to another audio amplifier (tube *No. 3*) and the rest of the signal is applied to the cathode of a diode rectifier. A diode rectifier works in such a manner that when the plate is made more positive than the cathode or the cathode made more negative than the plate, the tube will conduct, because

electrons will be emitted from the cathode and electron current will flow from the cathode to the plate *as long as the plate is positive with respect to the cathode*. The audio signal flowing toward the rectifier is varying continuously above and below its d-c level as any audio signal does.

Now when tube *No. 2* (the diode rectifier) does not receive any signal on either of its elements, no current flows in the tube because the difference in potential between the plate and the cathode is zero. When the audio signal flows toward the cathode of the rectifier, current will flow in the tube as shown (from cathode to plate) but *only* during the *negative* half-cycle of the audio signal. In other words, the negative half-cycles of the audio signal makes the potential on the cathode of the tube more negative with respect to the plate and electrons will therefore flow from the cathode. The rectified current will flow as shown and there will be a drop across the diode load resistor *R*. The polarities on the resistor *R* will be such that at the plate end, there will be a negative potential and a positive one on the ground end. Attached to the negative side of this resistor is the grid-leak resistor of tube *No. 3* and in this manner a bias is applied to the grid of the *No. 3* tube. It then follows that if a large audio signal is applied to the *No. 1* tube a large signal will likewise be applied to the cathode of *No. 2* tube. The *No. 2* tube, or rectifier, will then supply a greater negative bias to *No. 3* tube, thus reducing the amplification of the tube and in turn the output is
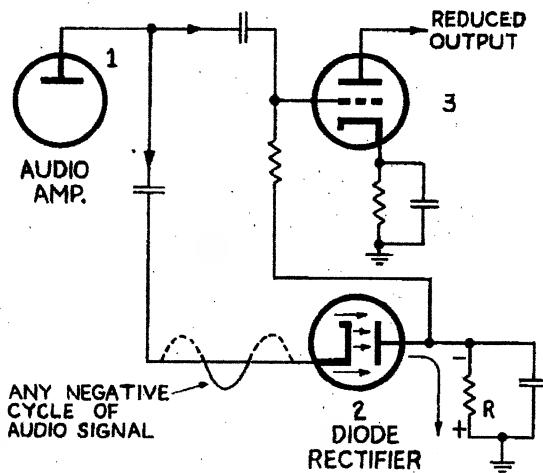

Fig. 3.—A simplified volume-compression circuit in which the audio gain is increased when the signal is weak and the gain reduced when the signal to be amplified is large.

also reduced. If the signal drops below a certain level, then no bias will be applied from the rectifier tube and tube *No. 3* will function normally according to its own self-bias. With this general analysis of how a typical volume-compression circuit works, let us now refer to the different recording positions of this model.

Radio Record Position

With the radio record push button depressed, most of the regular radio circuit is used in conjunction with the necessary recording equipment. In Fig. 4 is shown that section of the "radio-record" position that is of interest to us. Upon close examination of this circuit it will be noticed that no manual controls are included in this position. The control of the audio output in this circuit is accomplished by automatic methods. There is the regular avc control of the r-f tubes plus a volume-compression control in the audio circuit. To understand the function of the different tubes and component parts used in this position, it would be best to trace a signal through the circuit. Since the r-f section of this set is of the usual type, it is omitted and the rest of the circuit

that is used in the radio record position is shown.

After the r-f signal is converted into the i-f signal, it is impressed (through the secondary of the 2nd i-f transformer) across the diode section of *V6*, the 6SF7 tube, for detection and rectification. The rectified and detected signal is now at the junction between *R14-1* and *R13-2* after the proper r-f filtering. From this point, it is taken off and filtered through the audio filter, capacitor-resistor combination of *R1-2* and *C8-1*, to become the regular avc signal. Also from the above junction point between *R14-1* and *R13-2* the rectified and detected i-f signal, which is now the modulated envelope of the original r-f carrier input signal, is impressed as an audio signal across the complete capacitor-resistor combination consisting of *R13-2, C12, R12-2* and *R14-5*. This audio signal is then tapped off between the junction of *R12-2* and *C12* and it is impressed onto the control grid, pin 2 of the first audio tube

BAND-SWITCH SHOWN
AT 1ˢᵀ POSITION.
BROADCAST BAND
(RADIO RECORD)
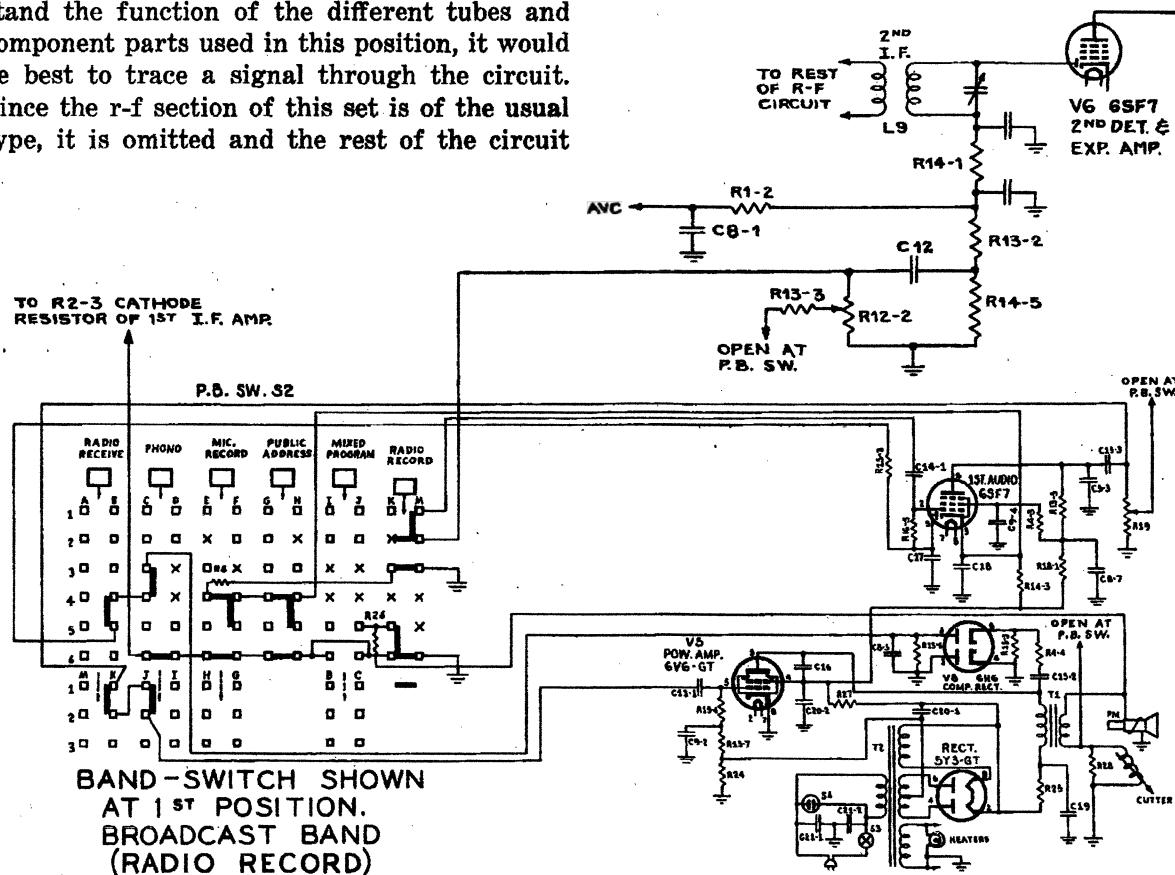
Fɪɢ. 4.—When the "Radio Record" push button is depressed in the Packard Bell Model 1052, most of the regular radio circuit is used in conjunction with the recording equipment.

(6SF7) through pin *M1* of the top of the push-button switch and also through capacitor *C14-1*. The audio signal is then amplified by the first audio tube and appears on the plate as an amplified signal. From the plate of this tube the audio signal goes through capacitor *C13-3* to the push button switch (lower part) through contacts *K1*, *K2*, *J1*, and *J2* then through capacitor *C13-1*, to be impressed onto the control grid circuit and pin 5 of the 6V6GT power amplifier. The signal is still further amplified by this last audio tube and it is applied directly to the high side of the primary of the output transformer *T1*.

Up until this point there was no evidence whatsoever that any of the audio signal was being controlled. From this point on it will be shown how volume compression in this circuit works to serve the purpose of an automatic *audio* control circuit.

The 6H6 *(V8)* tube serves as the compression rectifier. Only one diode section of this tube is used, namely pins 5 and 8, while the other diode section is inactive because pins 3 and 4 are grounded. For the sake of simplicity the compression circuit is extracted from Fig. 4 and it is redrawn in Fig. 5, only those components of interest in analyzing the control circuit being shown.

The Volume-Compression Circuit

What happens in this compression circuit is that some of the output audio signal is tapped off the plate of the 6V6GT, the same as the high side of the primary of the output transformer. It is then fed onto the cathode (pin 8) of the 6H6 compression rectifier. The amount of audio fed to the cathode is determined by the audio voltage dividing network of *C15-2*, *R4-4*, and *R15-5*. The total audio output is across the 6V6 plate to ground and therefore impressed across points *A* and *B* in Fig. 5. The audio voltage drop in the *R15-5* (470,000-ohm resistor) is that amount applied onto the compression rectifier cathode.

As previously described, when the *negative* swing of the audio signal is on the cathode, electrons will flow from the cathode to the plate. (The electron current will flow, however, *only* on the negative swing of the audio signal). Due to this current flow a rectified bias is developed across the diode plate (pin 5) resistor

R13-6 with potentials on the resistor as shown. Since the audio voltage is continually varying in amplitude, so is the drop across the *R13-6* (220,000-ohm) diode load resistor. Therefore the drop across *R13-6* is shown in Fig. 5 as a variable voltage. This voltage e_g, is applied to the control grid (pin 2) of the 6SF7 first audio tube through the high-resistance grid leak, *R16-5*. This bias, added to the self bias already existing on the tube, reduces the tube amplification and therefore its output is also reduced. It can then be stated that when the rectified bias on the first audio grid is increased, the first audio tube will reduce its output, and when this same bias is reduced, the output of the first audio tube increases.

Fundamentally what is happening is that the output of the first audio tube is being controlled by the rectified bias from the 6H6 compression rectifier. This simple control constitutes a method for keeping the audio level output constant. Another beneficial feature about this type of control is that it is completely automatic.

The audio volume is kept at a constant level in the following way: Assuming a gradual increase of audio voltage beyond what is desired, then the output from the 6SF7 first audio tube is also increased. This increased output constitutes a greater input to the 6V6GT power output tube, which in turn causes a greater output from the 6V6GT tube itself. This being the case, then a greater audio voltage is applied onto the cathode (pin 8) of the 6H6 compression rectifier than previously existed there. This makes the plate (pin 5) of the 6H6 tube much more positive with respect to the cathode but *only* during the negative half cycles of the audio signal. Since the difference in potential
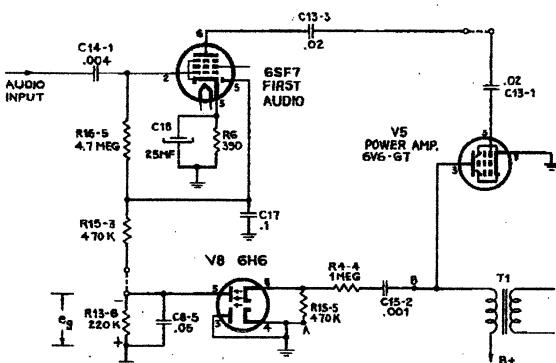

Fig. 5.—The simplified circuit of the volume compression portion of the Packard Bell Model 1052.

between these two diode elements is increased, the electrons flowing from the cathode to the plate are likewise increased and there is a larger rectified d-c voltage drop e_g across the diode plate-load resistor R13-6. The capacitor C8-5 (.05 mf) serves as an audio bypass capacitor for any audio frequencies that may appear in the plate circuit of the 6H6 compressor tube. The larger e_g drop that occurs puts a larger negative bias on the grid of the first audio tube and this increased bias reduces the output from this tube. The reduced output from this latter tube in turn causes a reduced input to the 6V6GT tube and naturally the output from this 6V6GT is also reduced. Since the output of this last tube is reduced, then the audio voltage applied to the cathode of the compression rectifier is also reduced, which means a reduced e_g drop, the application of a smaller negative bias on the first audio tube, and therefore an *increased* first audio output.
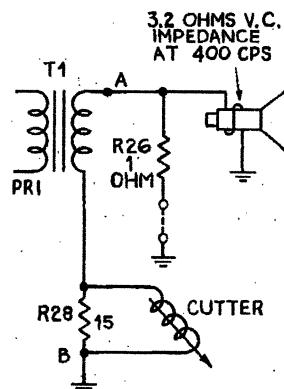

Fig. 6.—Simplified circuit of the loudspeaker and recording circuit of the Packard Bell Model 1052.

From the above analysis, it is consequently seen how the volume-compression circuit automatically controls the audio output by keeping this output at as constant a level as possible. With the automatic controlling of this circuit now considered, the next thing to do is to trace the audio signal from the 6V6GT plate.

Referring back to Fig. 4, the audio output from the plate of the 6V6GT tube is impressed across the primary of the output transformer T1. The *controlled* audio signal is then impressed across the secondary of T1 by transformer action and the total audio voltage now appears between the high side of the T1 secondary and ground. The circuit is redrawn in Fig. 6 in such a way as to make it easier to understand.

The total audio voltage in this circuit appears across points A and B. Since point B is ground,

then the total output exists between point A and all ground points common to this output circuit. In other words, some of the audio voltage is impressed across R26 in parallel with the voice coil and some appears across R28 in parallel with the cutter. The speaker, when used in the recording circuits, is used as an audible indicating device to inform the operator what type program he is going to record. The R26 (1-ohm) resistor across the voice coil is to reduce the effective impedance of this voice coil, which is 3.2 ohms, in such a manner that there will be a mismatch between the voice-coil circuit and the output transformer; consequently, the sound heard in the speaker will be slightly muted. Resistor R28 in parallel with the cutter is to insure proper matching purposes for the impedance that the cutting head offers. The compressor circuit assures the recording mechanism that on loud signals the audio volume will be automatically adjusted to a constant level and the needle of the cutter will be prevented from cutting into adjacent grooves.

Microphone Record

In the microphone recording position the same volume compression circuit also plays an important role in automatically controlling the audio volume cut into the record. It functions in exactly the same way it did in the radio record position. In Fig. 7 the circuit of the microphone recording position is shown. In this circuit no part of the regular radio section appears because there is not any need for r-f selection in this type of recording, since all that we are concerned with is recording *only* what sound goes into the microphone. In this circuit there is incorporated a 6SQ7 tube used as a microphone amplifier. Upon quick examination of the complete circuit, it will probably appear the same as the audio section of the previous recording position except for the addition of the microphone amplifier. Physically speaking, they do appear very much the same, but electrically they are somewhat different. These differences will be noted as a signal is traced coming from the microphone into the microphone amplifier.

Assuming some sound enters the microphone, then the 6SQ7 microphone amplifier tube increases the audio signal coming from the microphone and this newly amplified audio signal appears on the plate of the 6SQ7 tube. From

here the signal goes through C3-6, through the pins No. 2F, 1F, 1G, 1H, 1I, 1J, 1K, and 1M of the top section of the push button switch, through C14-1 to the control grid of the first audio 6SF7 tube. From this point on, the first audio tube, the 6V6GT power output tube, and the 6H6 compression rectifier function in exactly the same way as in the previous recording position. The only difference in the circuit is that instead of the 390-ohm cathode biasing resistor (R6) of the first audio tube, there is now a 270-ohm cathode bias resistor (R7) there (shown on the left of Fig. 7).

The reason for this resistor change is that in microphone recording there is need for a smaller self-bias on the first audio tube as compared to the self-bias needed for radio recording. The audio signal, after being impressed across the primary of the T1 power output transformer, is then impressed across the secondary of the T1 by transformer action. Since the operator of the microphone himself knows what will go into the record, there is no need for the loudspeaker to be used. It will therefore be noticed that the voice coil of the speaker is shorted to ground through pin Nos. 2G and 3G of the lower section of the push button switch. The only operative unit in the output circuit is the recording mechanism which functions the same as in the radio record position.

Mixed Program

In this unusual type of circuit, radio record and microphone record are combined in one complete circuit to allow a mixing of both types of recording. With it a person can, for instance, sing over the microphone to the music of some famous orchestra coming over the radio and it will appear as a completely single recording. In combining these two different types of recording circuits, a few interesting circuit changes occur that are worth noting. In Fig. 8, that part of the "mixed program" circuit of benefit to us is shown. The regular r-f section that is inserted in the mixed program circuit is the usual schematic and it is omitted for the same reasons it was done in the "radio-record" position. The most important part of this "mixed program" circuit is the beginning of the audio input sections where both the audio signal from the radio section of the set and from the microphone section are mixed together. In practically all types of audio circuits where two or more signals are to be combined, an "audio mixer circuit" is invariably used. In order to understand this "mixed program" circuit it will be best to trace signals from both audio sections of this position.

Referring to Fig. 8, again we find that V6, the 6SF7 tube, performs the function of second detection of the i-f signal from the radio portion of the set, and that the audio signal from the radio section is found to exist across the R12-2, R14-5, and C12 combination, as it did in the radio record position. However, as shown in Fig. 8, the audio signal is tapped off this combination by the variable arm of the poten-

BAND-SWITCH SHOWN
AT 1ST POSITION.
BROADCAST BAND
(MIC. RECORD)
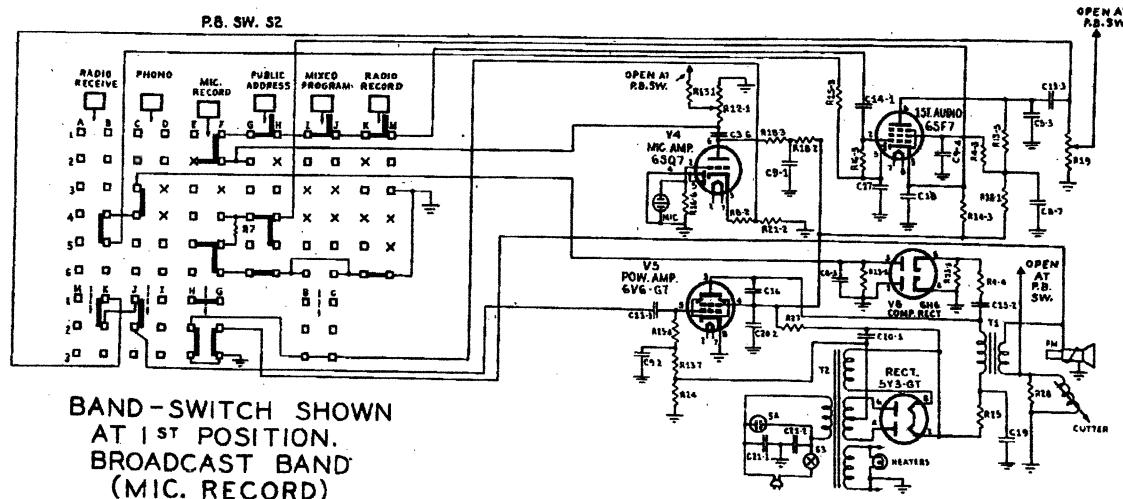
Fig. 7.—When the "Microphone Record" push button is depressed in the Packard Bell Model 1052, the radio broadcast portion of the set is switched out and only the microphone, the a-f amplifier with its volume-compression circuit, and cutter are functioning.

tiometer *R12-2*. Tracing the audio signal further it is found to travel through the resistor *R13-3*, to pin *2J* of the top section of the push button switch.

Stopping at this point for a moment let us now refer to the other audio system incorporated in this position, namely the microphone circuit. Looking at this section of Fig. 8, it is found that a signal from the microphone finds its way through the microphone amplifier (6SQ7) and it is then impressed across the *C3-6* and *R12-1* combination. The variable arm of the *R12-1* potentiometer taps off a certain portion of this microphone audio signal and the signal travels through *R13-1* to pin No. *2I* then to *2J* of the top section of the push button switch.

What we now have is both audio signals meeting at a common junction and if we trace *both* signals any further, it will be found that they are impressed together across the control grid of the 6SF7 first audio tube. This first audio tube functions as an "audio mixer" tube in this circuit. To make it easier to understand, the complete mixer circuit is redrawn and shown in Fig. 9. From this drawing, it is obvious how both separate audio signals meet and are impressed across the first audio control grid through *C14-1*. Any mixer circuit that combines their audio signals in a similar manner to the circuit of Fig. 9 is usually known as a "resistance-mixer" circuit. Of importance in such a circuit are the resistors *R13-3* and *R13-1*, both 220,000-ohms for this particular

set. The primary reason they are used is to prevent one potentiometer control from shorting out the other. In other words, if the variable arm of one potentiometer happened to be turned completely to ground, then without the 220,000-ohm resistors in the circuit, the arm that was turned to ground would short (to ground) the audio signal from the other potentiometer. Although the resistances *R12-1* and *R12-2* are variable, their variable arms are set to a predetermined value for the correct amount of audio signal mixing to give the operator of the set the desired level of audio that is to be cut onto the record without him making any adjustments. Since it may be necessary at some time or other to readjust these resistances, their variable arms are terminated in screwdriver adjustments. Coming back to Fig. 8 again, we find that from pin No. *2J* of the top section of the push button switch both audio signals are traced to pins No. *1J* to *1K* to *1M* then through *C14-1* to be
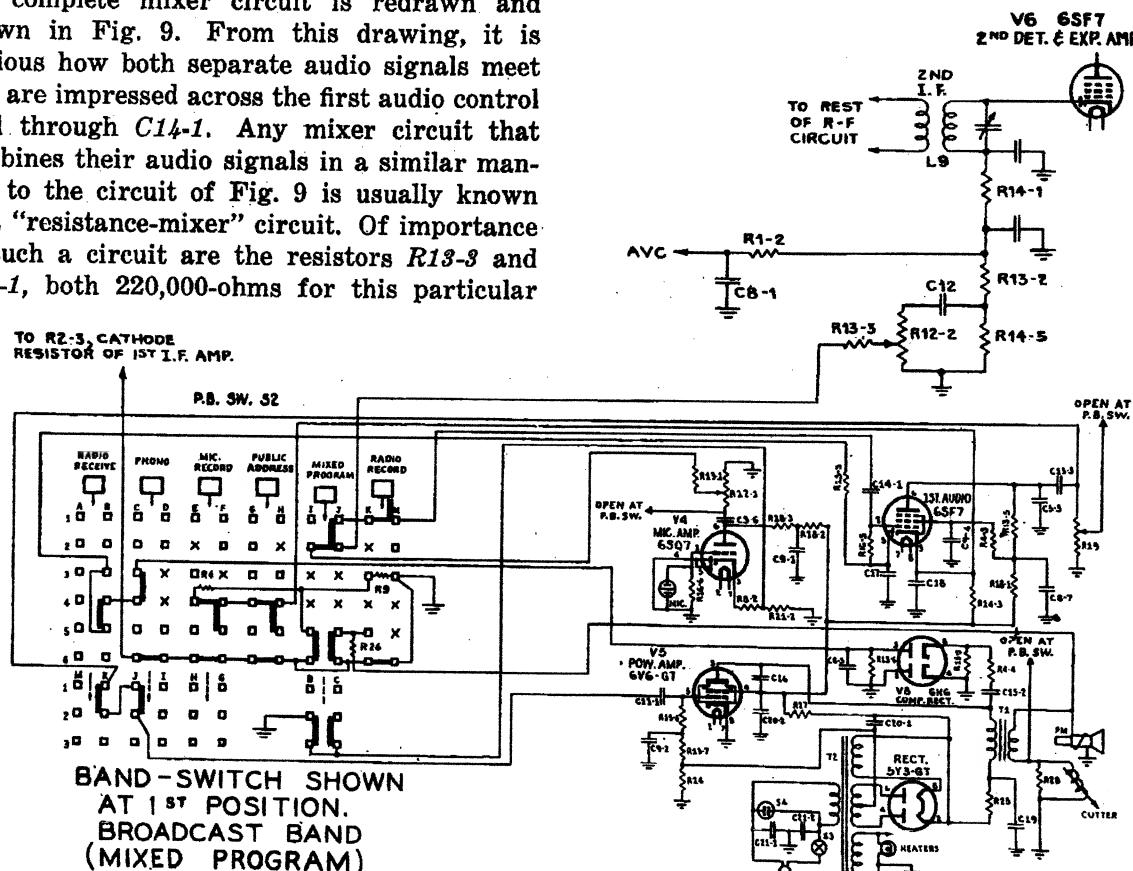

Fig. 8.—With the "Mixed Program" push button depressed in the Packard Bell Model 1052, the radio broadcast receiver and the microphone circuits are functioning.
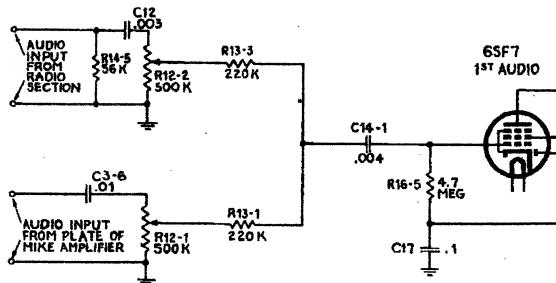
Fig. 9.—Schematic diagram of the mixer circuit of Fig. 8.

applied across the control grid of the first audio tube. This tube serves as an audio mixer and will not allow the output level of one signal to rise above the other and hence it will keep the output of all its audio signals at a constant level. In view of the fact that an overall gain control of the combined audio signals is needed, the volume compression circuit is still used in this position. In fact, the rest of this audio circuit, from the plate of the first audio tube, is the same as that of the "record-radio" position of Fig. 4, with one very slight difference: this is that the cathode bias resistance of the first audio tube is changed. In this cathode circuit there is now the 390-ohm resistor *R6*, in series with the 1200-ohm resistor *R9*, to effectively increase the bias on the first audio tube for the proper operation of the mixed programing. The volume compression circuit works in exactly the same way as previously described.

SPIEGEL MODEL TR-640

In the Spiegel Model TR-640, two different types of recording are incorporated—namely, radio recording and microphone recording. A few features of this recording set are different from those previously analyzed. The first major difference is in the method of controlling the amount of audio volume cut into the record that is being made. The volume control is used here on both recording positions to *determine* the amount of volume needed but in order to know where to set the volume control for the correct amount of audio, just plain *practice* is required. In other words, when the volume control is set to a certain position during recording and if it is found that the width of a groove on the record approximately equals the width of the uncut surface between the grooves then the setting is correct. Of course, this does sound a little crude in comparison to visual and automatic controlling, but it is nevertheless a good workable system when the operator of the set becomes familiar with it. The portion of the "radio recording" circuit of interest in this model is shown in Fig. 10A while that part of the "microphone recording" circuit is shown in Fig. 10B. Both schematics are presented together to show the similarity between their circuits.

Upon examining these schematics, it will be found that the speaker is inserted in both circuits and that a resistor *R10* (100-ohms) is between the high side of the voice coil and the high side of the secondary of the output transformer. This resistor slightly mutes the output of the speaker so as not to have the usual loud output interference with the recording. An un-
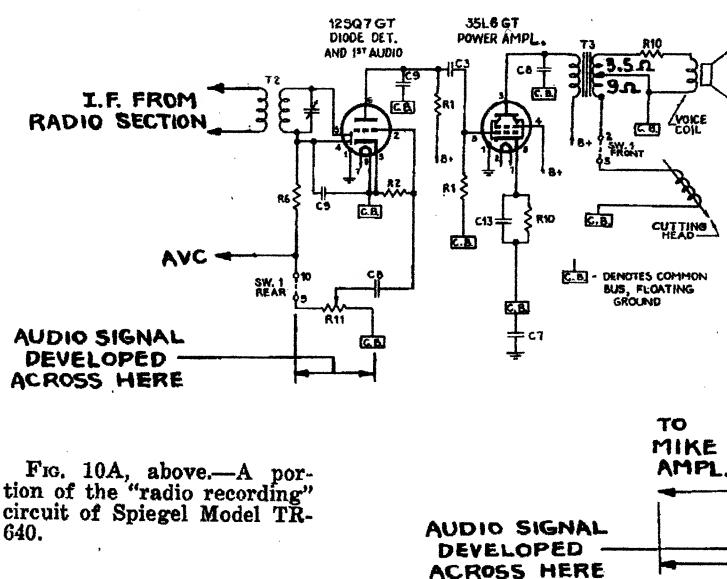

Fig. 10A, above.—A portion of the "radio recording" circuit of Spiegel Model TR-640.

Fig. 10B, below.—A portion of the "microphone recording" circuit of Spiegel Model TR-640.
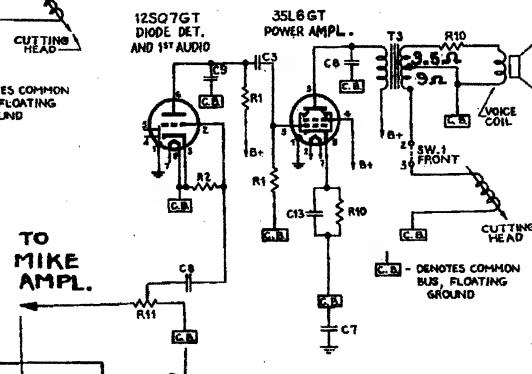
usual feature incorporated in this set is that in "microphone recording" the speaker is in the circuit the same as it is in "radio recording". The speaker is not really needed in this circuit but it can be used as a check to determine whether or not the audio signal is getting into the output circuit since there is no ready method of visual checking.

Another very interesting feature common to both types of recording circuits is that the output transformer has a split secondary. The upper half of the secondary serves the purpose of that section of the transformer used to match the impedance of the voice coil of the speaker. Since the complete audio output signal appears across the whole secondary coil, then part of this audio also appears across the lower half of the split secondary. The secondary is split in such a manner that an impedance of 3.5-ohms is in the top half and an impedance of 9-ohms in the lower half. The audio signal that appears across the lower half of the secondary is impressed directly across the cutting arm of the recording mechanism. The 9-ohm impedance of the lower half of the secondary is thus chosen so as to offer the proper impedance match for that impedance the cutting head itself offers.

BIASING METHODS

Bias for the proper operation of vacuum tubes is obtained by various methods. Bias is defined as the d-c voltage at the control grid of a tube measured with respect to the cathode, or to the filament in tubes having no cathode. The simplest, although not always the most desirable means of obtaining bias, is through the use of a battery. Fig. 1 shows how a 1.5-volt
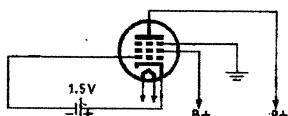

Fig. 1.—Fixed bias using a battery.

battery is connected between grid and cathode of a tube, with the negative terminal of the battery connected to the grid and the positive side of the battery to the cathode, thus providing a —1.5-volt bias.

With fixed d-c potentials on the plate and screen grid of a tube, the bias on the control grid of that tube will determine the operating point of that tube, that is, the amount of plate current which will flow. As the bias becomes more negative, more and more electrons within the tube are repelled by the grid from their normal flow to the plate until the grid reaches a point so negative that plate current ceases to flow, this point being known as "cut-off bias." Conversely, as the bias becomes more positive, more plate current will flow until it becomes excessive and the tube breaks down.

In order to obtain bias on the grid of a tube, it is not necessary to use an outside source of voltage, such as a battery, because this bias voltage can be developed from the current that is caused to flow from a tube. Fig. 2 illustrates such a method, which is called grid-leak bias.
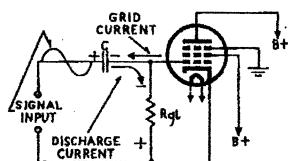

Fig. 2.—Bias attained by the use of a grid-leak resistor.

The bias is developed by the action of the components C, the grid capacitor, and Rgl, the grid-leak resistor. If the input signal, a sine wave in this case, is impressed between grid and cathode, on the positive swing of the input

signal, the grid becomes positive, causing grid current to flow in the direction of C, as shown by the arrow, and charging up this capacitor negatively on the grid side. Then as the input signal swings negative, grid current ceases to flow, but the negatively charged capacitor discharges through the grid-leak resistor making this resistor negative, on the grid-side, with respect to the cathode. This charging of the capacitor and discharge through the grid-leak, keeps a negative bias on the grid during the time that the signal is being applied.

Fig. 3 illustrates another means of obtaining a negative bias due to the normal current flow in a tube, without considering any input signal, that is, with the tube in a quiescent state. With the proper voltages on the screen grid and plate of the tube, current flows from

Fig. 3.—Bias attained by the use of a cathode resistor.
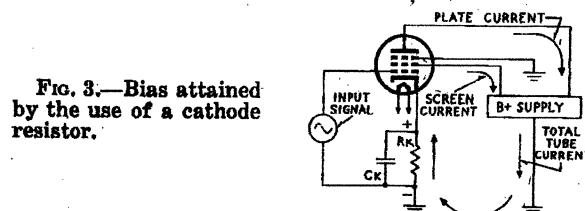
the cathode to the screen and plate, then to the supply-voltage source and then combining, forms the total tube current or as it usually is called, the cathode current. This current flows to ground; then from the grounded side of the cathode resistor, R_k, through this resistor and back again to the cathode. The current flowing through the cathode resistor R_k causes a voltage drop across the resistor with polarities as shown, that is, negative at the grounded side with respect to the cathode side. This voltage developed across the cathode resistor is the grid bias, and it is a negative bias, since bias voltage is measured from the grid to the cathode.

These are the three basic methods of obtaining bias: 1, the use of a bias battery cell; 2, grid-leak, and 3, cathode-resistor developed bias. How and why these methods and variations of these basic methods are used in current radio sets are discussed below.

Garod Model 6DPS8

Fig. 4A shows the i-f amplifer in the broadcast band of Garod Model 6DPS8. In order to obtain bias for this tube in the broadcast band, a 470-ohm resistor, which is bypassed by a 0.05-mf capacitor, is used in the cathode circuit. The total plate and screen current of the *6SK7* tube flows through this resistor, causing a voltage drop of about 3.5 volts across it. The
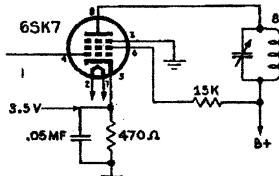
FIG. 4A.—Biasing of the i-f amplifier in the broadcast band of the Garod Model 6DPS8.

bias for this tube then, with no signal being received, is —3.5 volts, which gives sufficient gain with a good signal-to-noise ratio. The amount of current flowing through the cathode resistor can be calculated by Ohm's Law,

$I = \dfrac{E}{R}$, which in this case is $\dfrac{3.5 \text{ volts}}{470 \text{ ohms}}$ or 7.4 milliamperes.

When this set is switched to the short-wave band, increased gain is required at the higher frequencies to improve reception. In order to obtain more gain, a less negative bias is used
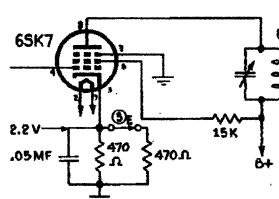
FIG. 4B.—Biasing of the i-f amplifier in the short-wave band of the Garod Model 6DPS8.

on the i-f amplifier grid, and the bias is reduced to about —2.2 volts. In order to obtain this less negative bias, the resistance was decreased in the cathode circuit. Fig. 4B shows that in the short-wave band, another 470-ohm resistor is put in parallel with the cathode network by closing switch *3E*. The equivalent resistance of the two 470-ohm resistors in parallel is 235 ohms. This makes the cathode current 9.4 milliamperes, an increase of about 2 mils, and the i-f stage is operated with greater gain.

Howard Model 920

Most oscillators use a grid-leak method for obtaining bias on the oscillator grid. This bias

voltage will vary, since the various frequencies of the oscillators will cause different amounts of grid current to flow through the grid-leak resistor. In the Howard Model 920, the grid-leak resistance Rgl in Fig. 5, in the converter stage, is actually composed of two resistors in series, one being 100,000 ohms and the other 68,000 ohms, making the total grid-leak resistance 168,000 ohms. The voltage at the oscillator grid, across the grid-leak will vary from —9.5 to —12 volts. It was found convenient to use a portion of this voltage, developed across the oscillator grid-leak resistance, for biasing the *3Q4* output tube. The way this was accomplished was by connecting the grid, pin 3 of the *3Q4* output tube through a 470,000-ohm isolating resistor to the junction on the
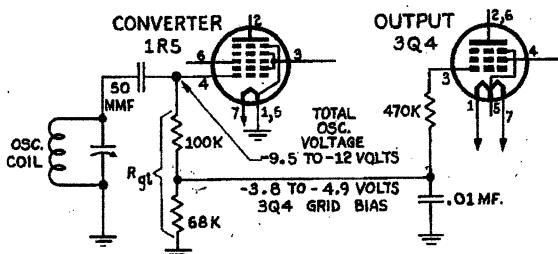
FIG. 5.—Grid-leak resistor used in the Howard Model 920.

grid-leak between the 100,000-ohm and the 68,000-ohm resistors. The bias for the *3Q4* tube will be that portion of the oscillator voltage which is developed across the 68,000-ohm resistor. This bias is $\dfrac{68}{168}$ of the total oscillator voltage and will vary from —3.8 to —4.9 volts.

Goodrich Model R643W, Stewart Warner Model 62T16, and Galvin Model 10T

Another means of obtaining the high bias required for output tubes can be accomplished by connecting the proper amount of resistance in the negative return of the power supply used. Figs. 6-A, -B, and -C, show three models using this system. In the Goodrich Model R643W, a 90-volt B battery is used for the supply voltage. As shown in Fig. 6A, the high side of the battery supplies the required voltage to the plates and screens of all the tubes in this set. From the negative side of the battery, a 390-ohm resistor, *R11* is connected to
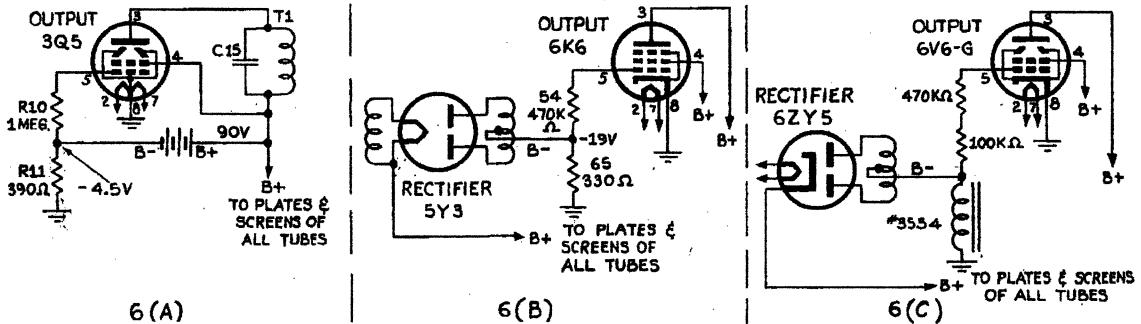
Fig. 6.—Grid bias obtained by connecting resistance in the negative return of power supply in the Goodrich Model R643W, Fig. 6A; the Stewart Warner Model 62T16, Fig. 6B; and the Galvin Model 10T, Fig. 6C.

ground. All the direct currents of the plates and screens of the tubes have to flow to their source of supply, the B battery, and since $R11$ completes the battery path to ground, then all these currents also will have to flow through $R11$, which causes a —4.5-volt drop across it. The grid of the $3Q5$ output tube is then connected through $R10$, a 1-megohm resistor, to $R11$ and the negative side of the battery. This negative voltage, —4.5 volts, is the bias applied to the grid, pin 5, of the output tube. The total current flowing through resistor $R11$ can be computed very simply by Ohm's Law,

$$I = \frac{E}{R}, \text{ which is } \frac{4.5 \text{ volts}}{390 \text{ ohms}}$$

giving 11.5 milliamperes.

In the Stewart-Warner Model 62T16, Fig. 6B, the bias for the output tube again is obtained from the voltage developed across a resistor in the negative return of the power supply, this resistor 65 being a 330-ohm wire-wound resistor. The B+ supply voltage in this case, instead of coming from a battery, is obtained from the $5Y3$ rectifier tube. The high side of this rectifier, that is, the B+ voltage supplied to the plates and screens of all the tubes in this set, is obtained from one of the filament leads. The low (negative) side of the supply voltage is the center tap of the plate transformer. Again all the d-c plate and screen currents will have to flow to their source of supply, the rectifier, to B— and through the 330-ohm resistor to ground. The current flowing through resistor 65 causes a 19-volt drop across it, that is, —19 volts to ground. The grid, pin 5, of the $6K6$ output tube being returned through a 470,000-ohm resistor, 54, to this —19 volt point, therefore, is utilizing this voltage as bias. The total current flowing

through resistor 65 can be calculated as before,

$$\frac{19 \text{ volts}}{330 \text{ ohms}} \text{ equalling } 57.6 \text{ ma.}$$

Fig. 6C shows how the bias is obtained on the $6V6$ output tube in the Galvin Model 10T. Here again the source of plate and screen voltage is a rectifier, a $6ZY5$ tube. The positive side is the cathode and the negative side is the center tap of the plate transformer, which is returned to ground through the choke, $No. 3554$. This choke is part of the rectifier filter network, but is wound to have just the proper resistance such that all the direct current flowing through it will give the proper voltage drop across it to be used as bias for the $6V6-G$ output tube.

Remler Model MP5-5-3

Automatic volume control is used to increase the gain of one or more tubes in a receiver when a weak signal is being received and to decrease the gain when a very strong signal is received. This is accomplished by putting a varying bias, depending upon the strength of the input signal, on the grids of the tubes whose gain is to be controlled. Fig. 7A shows the avc system of the Remler Model MP5-5-3.

When a signal is being received, an i-f signal will appear at the transformer $L-30233$, and this signal will be coupled through the 70-mmf capacitor to the diode plate, pin 5, of the $6SQ7$. When this plate is made positive (due to the positive swing of the input signal), with respect to the cathode, then the current will flow from the cathode to the plate, through the 3.3-megohm and 22-ohm resistors to ground. This will cause a voltage drop across these resistors, and it will be negative with respect to ground, the amount of this voltage depending upon the amount of current flowing through
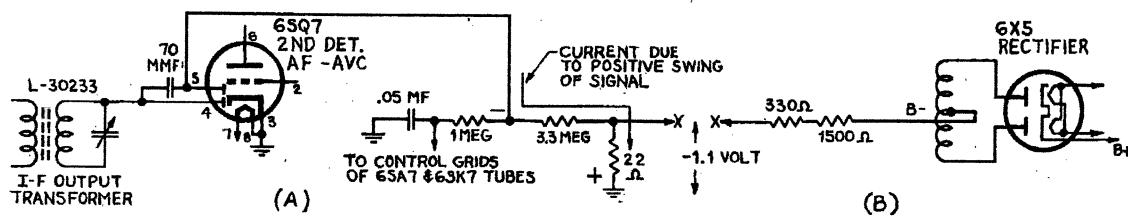
Fig. 7.—Remler Model MP5-5-3 having varying avc for increasing gain of the receiver.

these resistors. The current flow is directly proportional to the input signal, so that with a stronger signal, a greater voltage is developed across these resistors, and vice versa. This negative voltage is applied as avc bias through the 1-megohm and 0.05-mf filter, to the grids of the converter and i-f tubes. When no signal is being received, no avc voltage will be available for bias on these tubes, and without this bias, the plate current of these tubes may become excessive.

In order to prevent this excessive plate current, a minimum or residual bias must be supplied to the avc-controlled tubes from some other source when no signal is being received. In this receiver, the residual bias is obtained from a portion of the voltage developed across the negative return of the rectifier power supply, as shown in Fig. 7 at B. As stated previously, all the direct current flowing through the power supply, with no signal being received, will cause a voltage to be developed across the negative return of a power supply. The negative return here is composed of the 1500-ohm, 330-ohm, and 22-ohm resistors in series. (Points marked X in Fig. 7 actually are connected as one point). The currents flowing through this negative return causes —1.1 volts to be dropped across the 22-ohm resistor to ground. This bias of —1.1 volts is then fed through the avc filter to the grids of the 6SA7 and 6SK7 tubes and thus provides

Fig. 8A.—RCA Model 56X5 using residual bias on avc control tubes.

this voltage as minimum bias when no signal is being received.

RCA Model 56X-5

Another method of obtaining residual bias for avc-controlled tubes when no signal is being received, in order to protect the tubes from drawing excessive current, is to use a portion of the voltage developed across the oscillator grid-leak resistor as minimum bias for these tubes. Fig. 8A shows such a system as used in the RCA Model 56X-5.

Fig. 8B.—Equivalent circuit of residual bias arrangement used in RCA Model 56X5.

Here the avc voltage is developed due to rectification of the i-f signal at the detector diode plate, pin 4, across its load $R7$, which is 500,000 ohms. This negative avc voltage that is developed across $R7$ is fed to the r-f converter, and i-f control grids, through the avc filter consisting of $R6$ and $C14$. When no signal is received, the oscillator will still function and cause a voltage to be developed across its grid-leak resistor, $R4$, this voltage varying from —6 to —10 volts. This same voltage will also appear across the three resistors $R5$, $R6$ and $R7$ in series, which are in parallel with $R4$, as shown in Fig. 8B.

The voltage that is developed across the resistors $R6$ and $R7$ will be the residual bias used for the avc-controlled tubes when no signal is received. This voltage will vary as the total voltage, according to the proportion of the resistance of $R6$ and $R7$ to the total resistance of $R5$, $R6$, and $R7$, this being 3.8/18.8, which is approximately 1/5. The voltage will then be 1/5 of —6 to —10 volts or —1.2 to —2 volts, which is the residual bias for the avc-controlled tubes when no avc bias is available.

NEUTRALIZATION

Neutralization may be defined as the method of cancelling the effect of voltages fed back (usually through interelectrode capacitance coupling), to the input stage of an amplifier tube, by returning an equal voltage having opposite phase to the same input point through an auxiliary circuit. This section will discuss neutralization in converters and mixer-oscillator circuits, with the neutralization required for space-charge coupling, taken up in detail below.

For any stated value of cathode temperature, the density of space charge, created by the electrons in the space around the cathode, may be considered in equilibrium. In other words, for any given value of cathode temperature, which represents a certain rate of electron emission, a state of equilibrium (known as "emission saturation") is established whereby the maximum density of the space charge is fixed. The fact that electrons are being emitted continuously by the cathode does not increase this space-charge density, because the negative electrons emitted from the cathode are repelled back to it by the negative space-charge field. To express this differently, for every electron which gets into the space-charge zone after this state of balance or equilibrium has been reached, the space-charge field returns an electron to the cathode.

For every value of cathode temperature, a certain magnitude of thermal agitation occurs within the cathode-emitting material, and also a definite average initial speed for the emitted electrons. The higher the temperature of the cathode, the greater is the internal thermal agitation and the higher is the average speed of the emitted electrons. Equilibrium of the space charge at any cathode temperature is reached when the number of electrons which have formed the space charge is such that the magnitude of the repelling field (caused by the space charge) acting at the cathode is sufficient to overcome the speed of the emitted electrons and repel them back to the cathode.

If this state of equilibrium of the space charge is disturbed by the passage to the plate of an electron from the end of the space-charge zone nearest the plate, a reduction in the total magnitude of the field of the space charge will take place. The ability of the space-charge field to repel the electrons emitted from the cathode is reduced, permitting that number of electrons required to restore equilibrium to be added to the space charge in the zone near the cathode. More than the required number of electrons to replace those removed will not be added to the space charge, for the state of equilibrium previously mentioned is maintained automatically. It is possible to visualize a condition in which electrons are removed from one side of the space charge at a definite rate and electrons are admitted into the space charge at the same rate on the other side. Thus, there is a progressive movement of electrons from one side of the space charge, through it, and out of the space charge on the other side, without altering the density of the space charge.

It may thus be said that in an ordinary amplifier tube, the space-charge density remains fairly static once equilibrium is reached due to stabilization of cathode temperature and electrode d-c voltages.

Interelectrode Capacitances

It is well known that two metal plates placed in proximity form a capacitor. Fig. 1 schematically illustrates the electrode structure of a common type of converter tube. In this tube,

Fig. 1.—Schematic representation of a converter tube showing the positions of the various grids, plate, and cathode, between which there are capacitive effects.

CONVERTER
PLATE
SUPPRESSOR
SCREEN
SIGNAL GRID
OSCILLATOR ANODE
OSCILLATOR GRID
CATHODE
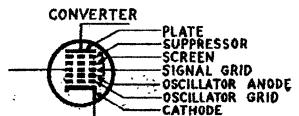
the oscillator grid forms one plate of a capacitor with the cathode as the other plate, and the oscillator grid also forms one plate of another capacitor with the oscillator anode grid as the other plate of this capacitor. Further, the oscillator grid and the signal grid also form a capacitor since the oscillator anode grid is not a perfect screen (electrostatic shield) and does not prevent many electrons from passing through its grid mesh and affecting the signal-grid potentials. Thus the space charge dis-

cussed above extends from the cathode through the oscillator grid and other grids to the plate. Since the distribution of the space charge is not uniform throughout this zone but varies depending on electrode potentials, tube geometry, and other factors, a portion of this space charge exists between the oscillator grid and the signal grid.

Since the space charge is a negative field, it exerts an influence upon neighboring electrodes making them "appear" more negative than they are. Thus with a static potential existing between two electrodes, the space charge has a tendency to reduce this potential and make it more negative or less positive than it is. Because the electrons are bunched together when passing through some grid openings (due to negative potentials on some of the grids), they exert a strong negative effect, so that fewer electrons are attracted to or impinge on the grid, and the current flow in the grid structure is thus decreased.

The potential of the oscillator grid (in a converter tube) varies from instant to instant at a rate of oscillation determined by the total oscillator circuit. This variation in the potential of the oscillator grid has the effect of varying the space charge from instant to instant, since the variation of oscillator grid voltage in a plus and minus direction about its fixed negative bias point, accelerates or retards the flow of electrons from the cathode at the rate of the oscillator frequency and so varies the space charge at the oscillator frequency rate.

Space Charge Coupling

The effect of this varying space charge on the signal grid is to induce a current at the oscillator frequency in the signal-grid circuit. Thus it may be said that the space charge couples the oscillator grid and signal grid through its electrostatic field.

According to the impedance of the signal-grid circuit, a voltage will be flowing at the oscillator frequency in the signal-grid circuit. If the intermediate frequency of the receiver is low, then the oscillator frequency, differing from the frequency to which the signal resonant circuits are tuned by the small value of intermediate frequency, will be presented by the signal circuit with a comparatively high impedance. This follows from the principles of parallel-tuned circuits. Since the signal circuit will thus be very selective for the oscillator frequency, a high oscillator current will flow in the signal-grid circuit. The high IZ drop (current times impedance) in the signal circuit means that a high oscillator voltage will be present in the signal circuit.

This feeding of the oscillator energy into the signal circuit is a disadvantage since it results in reduced tube-conversion gain. In addition, the tube-conversion gain will vary as the receiver is tuned to different frequencies. Tube-conversion gain is a function of oscillator voltage (up to the point of peak gain), so that when the oscillator voltage is reduced, the tube-conversion gain is decreased. The decrease in oscillator voltage as a result of this space charge coupling may be analyzed as follows:

As explained previously, there is a relatively large voltage drop at the oscillator frequency across the signal-grid circuit impedance. Most broadcast receivers are so designed that the oscillator frequency is higher than the signal frequency, so that the signal circuit offers a capacitive reactance to the oscillator voltage. This is due to the fact that a parallel-tuned circuit offers capacitive reactance to frequencies above its resonant frequency. Since the oscillator voltage is usually much greater than the signal voltages, this capacitive reactance is of some importance. The interelectrode capacitance between the oscillator grid and the signal grid due to space-charge coupling, in series with the effective capacitance that the signal circuit presents to the oscillator frequency, may be considered to be in parallel with the interelectrode capacitance between the oscillator grid and cathode. These three capacitances, the interelectrode capacitance between cathode and oscillator grid C_{ok}; the signal-grid-to-oscillator-grid interelectrode capacitance due to space-charge coupling C_{sc}; and the effective capacitive appearance of the signal circuit to the oscillator frequency C_e, are shown schematically in Fig. 2. (Incidentally, there is a direct interelectrode capacitance between the oscillator and signal grids, as mentioned previously, thus shunting the capacitance due to space-charge coupling, but this is ordinarily very small, being of the order of 0.2 mmf in many common converter tubes, and so can be neglected in this discussion.)

It can be seen in Fig. 3 that the effective capacitance C_e' of the signal circuit for the oscillator frequency in series with the interelectrode capacitance due to space-charge coupling C_{sc}, shunt the oscillator tank circuit and the interelectrode capacitance between cathode and oscillator grid C_{ok}. Since the effective impedance of the two series capacitances is much less than the oscillator input grid impedance of the tube (effectively due to C_{ok}),
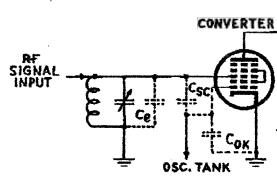

Fig. 2.—Various input capacitive effects in a converter tube: C_{ok}, between the oscillator grid and cathode; C_{sc}, between the oscillator grid and signal grid; and C_e between the signal grid and ground.

the oscillator thus is loaded down and the possible output voltage is decreased. As stated previously, since the tube-conversion gain is dependent on the oscillator voltage, the conversion gain will be decreased as a result of this loading effect. In addition, since this loading effect will vary as the value of the effective capacitance across the signal circuit (due to impressed oscillator voltage) varies, the tube-conversion gain will differ for different signals received. This may be seen from the fact that as the receiver is tuned to different frequencies, the ratio between the oscillator and signal fre-
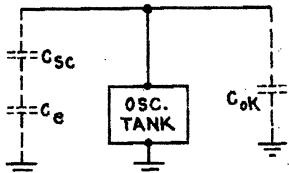

Fig. 3.—The loading effect of the three capacitances of Fig. 2 on the tank circuit.

quencies changes and the value of the effective capacitance across the signal circuit (due to impressed oscillator voltage) varies. Changes in the effective capacitance, C_e, result in different loading effects on the oscillator and thus the variation of the tube-conversion gain.

The energy transfer from the oscillator circuit to the signal circuits through the space-charge coupling and its sometimes disadvantageous effect of variation of tube-conversion gain, can be eliminated largely by a neutralizing circuit. This means that a voltage equal in amplitude to the oscillator voltage impressed across the signal circuit, but out-of-phase with it, is fed back from the oscillator circuit to the

signal circuit so as to cancel the above impressed oscillator voltage in the signal circuit. Usually, this is done by connecting a capacitor between the oscillator-grid circuit and the signal-grid circuit.

General Electric Model X-181 Series

The General Electric Model X-181 series receiver has this problem of space-charge coupling discussed above and neutralizes its effect by means of a fixed capacitor connected between the oscillator and the signal circuits. Fig. 4 shows a 1.0-mmf capacitor, $C35$, connected from the high side of the oscillator tuning capacitor to the high side of the mixer tuning capacitor. This capacitor serves to couple back this neutralizing voltage. The
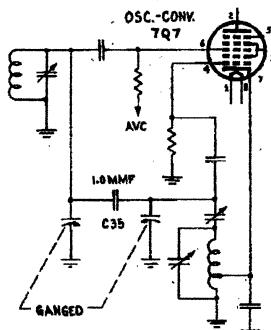
Fig. 4.—The capacitor $C35$ serves to couple back the neutralizing voltage from the oscillator circuit to the signal circuit.

same problem and solution can be seen in the Olympic Models 6-601W, 6-601V and 6-602 (Olympic page 15-5), and in other receivers.

Montgomery Ward Model 54WG-2500A

It must not be supposed that space-charge coupling between the oscillator and signal circuits can take place only where the two circuits make use of the same tube envelope. Where separate oscillator and mixer tubes are used and where the oscillator signal is fed into either the cathode or a different grid than the signal grid, space-charge coupling takes place between the oscillator and signal circuits. This coupling is ordinarily not great enough to require neutralization where separate mixer and oscillator tubes are used, whereas in the converter tube the spacing between the oscillator and signal grids is so small that the energy transfer from oscillator to signal grids is of importance. However in the Montgomery Ward receiver Model

54WG-2500A, where the oscillator signal is fed from the oscillator cathode to the mixer cathode as shown in Fig. 5, this space-charge coupling takes place.

The neutralizing signal, fed from the oscillator plate so as to be out-of-phase with the
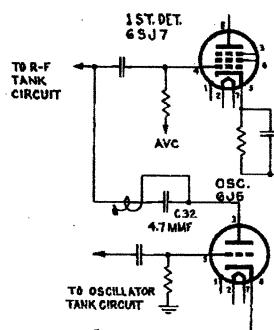
Fig. 5.—The neutralizing signal is fed from the oscillator plate through C32 and a gimmick, thus canceling the oscillator voltage appearing across the mixer input circuit due to space-charge coupling.

directly coupled signal, is fed through a 4.7-mmf fixed capacitor C32 and a gimmick, shunting this capacitor to the high side of the mixer input resonant circuit. This neutralizing voltage cancels the oscillator voltage appearing across the mixer input resonant circuit due to space-charge coupling. The gimmick shunting the neutralizing capacitor increases this neutralizing coupling to the required point. This latter point is explained more fully in the section on "Gimmicks."

Montgomery Ward Model 93WG-800

A similar problem of capacitance coupling not due to space-charge coupling effects, is encountered in the Montgomery Ward Model 93WG-800. In this receiver, because of various circuits that could not be eliminated, capacitance coupling takes place and oscillator voltage appears on the grid of the r-f preselector tube. As shown in Fig. 6 this oscillator voltage is neutralized by feeding an out-of-phase signal to the r-f grid by means of a coil winding and gimmick circuit coupled from the oscillator cathode coil to the r-f tube grid.

It is of interest to note that this neutralizing circuit is only effective on the short-wave band. (Montgomery Ward page 15-99 shows the neutralizing circuit included in the short-wave *clarified schematic* but not in the broad-

cast-band *clarified schematic.*) Since the neutralizing-circuit pickup coil is coupled to the short-wave oscillator coil, and the gimmick is coupled to the short-wave r-f coil, this coil-gimmick circuit cannot pick up energy from the broadcast-oscillator coil during broadcast operation and feed it to the broadcast r-f coil.

The reason for neutralizing on the short-wave band and not on the broadcast band may be explained as follows:

The ratio of the r-f signal frequency to the oscillator frequency is much lower on the broadcast band than it is on the short-wave band. For example, if the receiver intermediate frequency is 460 kc and the receiver is tuned to 1000 kc on the broadcast band, then the oscillator frequency is 1460 kc, and the ratio of r-f signal frequency to oscillator frequency is 68 per cent; on the short-wave band, assuming the receiver is tuned to 6000 kc, then the oscillator frequency is 6460 kc and the ratio of r-f
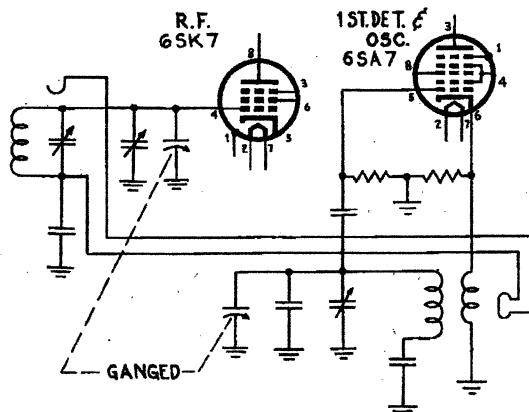
Fig. 6.—The oscillator voltage is neutralized by feeding an out-of-phase signal to the r-f grid by means of a coil and gimmick circuit coupled from the oscillator cathode coil to the r-f grid.

signal frequency to oscillator frequency is 92 per cent. Clearly, the oscillator frequency is much closer in terms of percentage to the r-f signal frequency on the short-wave band than it is on the broadcast band and as a result, the possibility of interaction is much greater on the short-wave band. For this reason, neutralization often is applied on the short-wave bands and not on the broadcast band in the same receiver.

PARASITIC RESISTORS

A parasitic resistor is one of the most common and effective methods of suppressing unwanted oscillations. These unwanted oscillations are called parasitic or spurious oscillations and may occur at either a lower or higher frequency than the desired frequency.

The effect of parasitics in a radio receiver will result in added plate dissipation, distortion due to overloading, and loss of power output due to the fact that this parasitic circuit will absorb the normal and useful energy from its associated circuit in order to sustain its own oscillations.

It is impossible to predict where parasitic oscillations will occur in a radio circuit. They are most commonly found in the local oscillator and amplifier stages and usually are due to the inductance of the leads in the plate and grid circuits, being in parallel with the interelectrode capacitance in the tube, that is, the grid to cathode (C_{gk}) and the plate to cathode, (C_{pk}) capacitances. In other words, feedback and resonance are caused between grid and plate circuits. This type of parasitic circuit will be of a higher frequency than the desired one. Low-frequency parasitics sometimes are produced by chokes of similar size located in grid and plate circuits. Fig. 1 shows
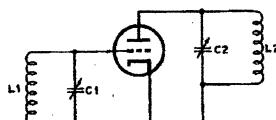
Fig. 1.—Typical circuit having a tuned-grid and a tuned-plate tank circuit.

a typical circuit with a tuned-grid and tuned-plate tank circuit.* This and similar circuits may have parasitic oscillations at high frequencies, for at these high frequencies the inductive reactance, since it varies directly with frequency, becomes very large and therefore $L1$ and $L2$ will act as an open circuit whereas the capacitive reactance, varying inversely with frequency, becomes very small, thereby having $C1$ and $C2$ act as a short circuit to these high frequencies. Fig. 2, shows the equivalent high-frequency parasitic circuit developed from the circuit given in Fig. 1, with the grid-cathode

*For an explanation of a tank circuit see the section on oscillators.

interelectrode capacitance C_{gk} in parallel with the inductance of the lead, from the grid through the short-circuited capacitor $C1$ and the plate-cathode interelectrode capacitance C_{pk} in parallel with the inductance of the lead from the plate through the short-circuited capacitor $C2$. This parasitic circuit is of the
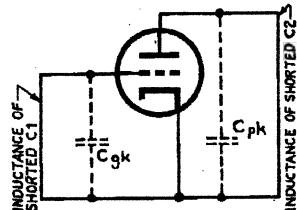
Fig. 2.—Resultant parasitic circuit of Fig. 1.

tuned-grid, tuned-plate variety, which is easily seen in Fig. 2 and the frequency of oscillation is higher than that desired since both the inductance and capacity of the parasitic tuned circuit are less than the values of the original circuit.

Fig. 3 shows a typical circuit where parasitic oscillations with a frequency lower than that
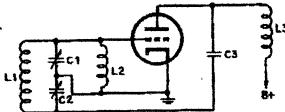
Fig. 3.—Typical circuit leading to low-frequency parasitic oscillations.

desired might occur. $L1$ is the tuning inductance, $C1$ and $C2$ are the tuning capacitances, $L2$ and $L3$ are r-f chokes. At low frequencies, the inductive reactance of $L1$ is negligible, but that of $L2$ and $L3$ becomes appreciable. The capacitive reactances of the tuning capacitors are also appreciable. Considering the above facts, it is possible to have a low-frequency parasitic circuit as shown in Fig. 4,
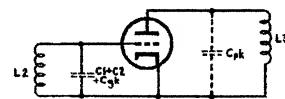
Fig. 4.—Resultant parasitic circuit of Fig. 3.

which is a tuned-grid, tuned-plate circuit where $L2$ in parallel with $C1$ plus $C2$ plus the grid-cathode interelectrode capacitance forms the grid circuit and $L3$ in parallel with the plate-cathode interelectrode capacitance C_{pk} forms the plate tuned circuit.

In order to damp these parasitic oscillations, comparatively small values of resistance, that is, from about 10 ohms up to about 150 ohms, are inserted in series with the grid and/or the plate leads thus reducing the effect of the parasitic oscillations and effectively leaving the original circuit unchanged. As will be shown, the grid and plate circuits are not the only places where parasitic resistors are used in radio sets nor is 150 ohms the largest value of parasitic resistor used, for in some cases resistors up to about 1000 ohms are found necessary to suppress the spurious oscillations. Usually one can identify a resistor as being a parasitic suppressor by its value and lack of any other function in the circuit.

Bendix Model 736B

In the Bendix Model 736B, Fig. 5, *R28*, a 10-ohm resistor, is placed directly in series with the signal grid, pin 8, of the converter, to suppress any parasitic oscillations occurring
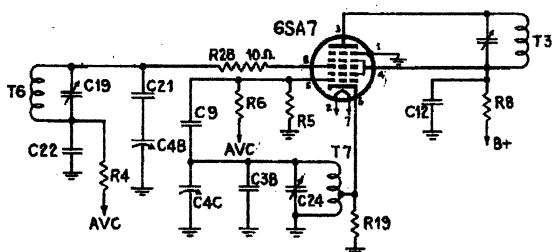
FIG. 5.—Bendix Model 736B converter stage.
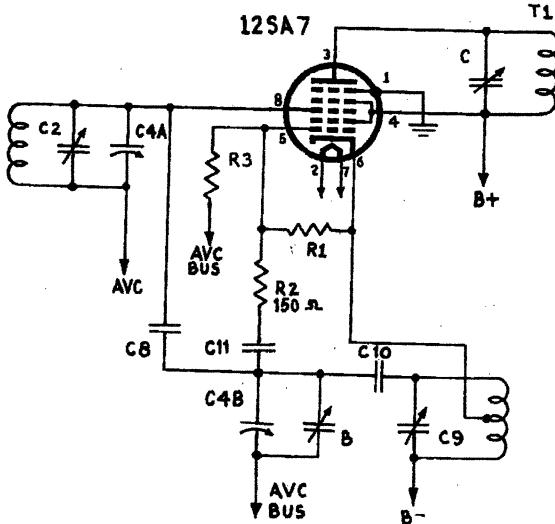
FIG. 6.—Converter stage of the International Detrola Model 568.

there. Note that this resistor is placed adjacent to the grid and outside of the tank circuit, composed of *T6*, *C19*, *C21*, and *C4B*, designed to produce the required resonant frequency.

Detrola Model 568

In the International Detrola Model 568, Fig. 6, there is a slight variation from the Bendix model. In this case, the parasitic circuit is again found in the converter stage in which a 12SA7 tube is employed. The electrical characteristics of this tube are the same as that of the 6SA7 tube, which was the one under discussion in the Bendix case. Due to the nature of the circuit, however, it was found that parasitic oscillations instead of occurring at the signal grid, pin 8, are here found at the oscillator grid, pin 5; therefore, the parasitic resistor, *R2*, which is 150-ohms, is placed directly in series with this grid, outside of the oscillator tank circuit so as not to change the oscillator frequency.

Pilot Model T-511

Fig. 7 shows the converter stage of the Pilot Model T-511. Here again a 12SA7 tube is being used, but the spurious oscillations that oc-
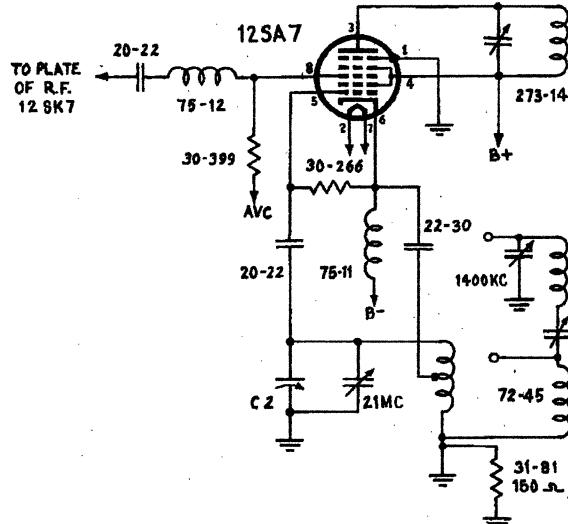
FIG. 7.—Converter stage of the Pilot Model T-511.

cur are not where one might expect them to be. Looking closely at resistor *31-81*, which is at the low side of the oscillator tank circuit, it appears as if this 150-ohm resistor is shorted since it is grounded at both ends, and there-

fore has no function. Actually, this resistor is the parasitic suppressor damping out oscillations that occur between stray capacities, components of the oscillator circuit, and inductance of leads between different potentials of the chassis. Although we usually think of a chassis as being at one potential, nevertheless at high frequencies, components of a circuit that are grounded at different points on a chassis may be at different potentials due to the distance and, therefore, the inductance between these chassis points will be different. In the Pilot model, the coil *72-45* is grounded to the chassis through a lead approximately 2 inches long. At the bottom of this coil is a lug to which the resistor *31-81* is soldered. The other end of this resistor is returned to a different point on the chassis, namely, the ground side of the 1400-kc trimmer and thus suppresses any parasitics that may occur between these various points.

Zenith Model 9HO79

These few examples show where parasitic resistors were needed in various portions of converter stages, in which it is rather common to have parasitic oscillations. The following will illustrate where resistors are used to damp spurious oscillations in amplifier stages of current radio models. Fig. 8 shows a 6K6GT tube
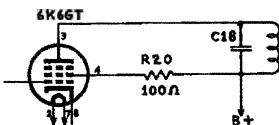
Fig. 8.—Zenith Model 9HO79 power amplifier.

being used as a single, power amplifier in the Zenith Model 9HO79. Parasitic oscillations, that might occur in this circuit without the use of *R20*, a 100-ohm resistor, to squelch them, would exist between the plate and screen grid of the tube across the output transformer and probably would cause distortion and hum in the output. In order to overcome the effects of these unwanted oscillations, it was found necessary to put the parasitic resistor directly in series with the screen grid, pin 4, rather than the plate where it is more commonly used.

G. E. Model X-181

Another fairly common source of trouble with spurious oscillations is found in tubes that are connected in push-pull, as shown in Fig. 9, in which is shown the output stage of

the General Electric Model X-181. The cause for parasitics in circuits of this kind can usually be traced to the inductance of leads of
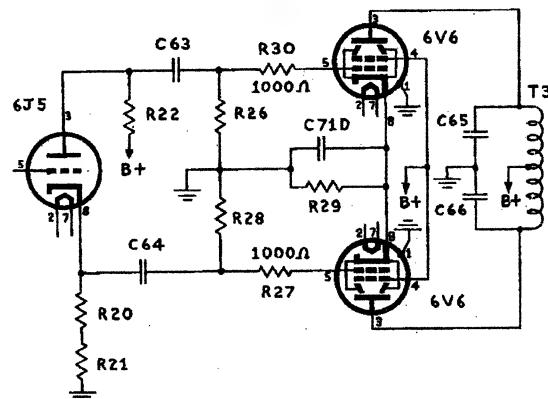
Fig. 9.—6V6 tubes in the push-pull stage of the General Electric Model X-181.

symmetrical wiring, in parallel with the interelectrode capacities of the tubes. This will then cause a tank circuit resonant to a high frequency and hence we have our unwanted oscillations. *R27* and *R30* are the parasitic suppressors used in this case. They are placed directly in series with the grids of both 6V6 tubes and note that these resistor values are rather high, being 1000 ohms each. The oscillations probably were of fairly great amplitude to require 1000-ohm resistors to damp these parasitics.

Westinghouse Model H-104

Parasitic oscillations are not always overcome or sufficiently damped out by putting resistors in the plate or grid circuits, for some-
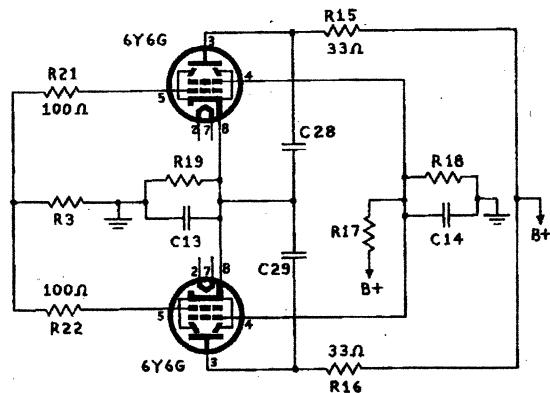
Fig. 10.—6Y6G tubes in the parallel output stage of the Westinghouse Model H-104.

times it becomes necessary to insert parasitic resistors in both the plate and the grid circuits. In Fig. 10 is shown that such is the case in the Westinghouse Model H-104. Here two 6Y6G tubes are connected in parallel in the output stage. When tubes are connected in parallel, if parasitic oscillations do occur, they are very frequently of the tuned-plate, tuned-grid variety producing oscillations of a high frequency. In order to overcome the effects of the spurious oscillations in the Westinghouse Model, a 33-ohm resistor was placed in series with each plate, that is *R15* and *R16*, and a 100-ohm resistor was also placed directly in series with each grid, that is, *R21* and *R22*.

In connection with parasitic oscillations and their cures, it is interesting to bear the following in mind, especially in small ac-dc receivers. Changes in lead dress, that is, departures created during service operations may give rise to magnetic coupling between points, which in turn may cause unwanted oscillations. Curing such faults with parasitic resistors is not recommended, it being more advisable to eliminate the unwanted magnetic coupling by re-dressing leads.

On the other hand, replacement of components with others of widely different characteristics may give rise to parasitics which did not exist before. When such oscillations develop, recall the nature of the service operation, the circuits involved, and the components replaced. In most cases, it will point to the possible location of the fault.

The location of parasitics in audio sections of receivers requires a logical method of operation. For that matter, the same is true regardless of where the parasitic may be located. They should be sought without any excitation (signal input) applied. Very often they may be at frequencies very far removed from those expected in the system. Inasmuch as systems are more sensitive with reduced bias, parasitics in audio circuits should be tested with the minimum safest grid bias, or zero bias if the plate and screen voltages can be reduced to safe values.

The cathode-ray oscilloscope is by far the best method of tracing parasitics in audio systems, because it will disclose these spurious signals over a very wide range of frequencies— especially when they are beyond audible range.

GENERAL DISCUSSION OF
SUPERHETERODYNE OSCILLATORS

The purpose of the oscillator in a superheterodyne receiver is to provide a signal which will beat with an incoming broadcast signal and produce in a mixer tube an intermediate-frequency signal which is similar in modulation characteristics to that of the broadcast signal. In some sets, the oscillator signal is developed by a separate tube and circuit which is coupled to a mixer tube; in others, the functions of the oscillator and mixer are combined within a single tube.

The arrangement employed to produce the i-f signal is shown in Fig. 1. This block diagram represents equivalent sections of a superheterodyne receiver which is tuned to a broadcast signal of 1000-kc. The 1000-kc signal is amplified by the r-f stage and applied to the input circuit of the mixer stage. The set oscillator supplies a 1450-kc signal which combines with the 1000-kc incoming broadcast signal in the mixer to produce an intermediate frequency which represents the difference between 1450 kc and 1000 kc or 450 kc. The i-f amplifier is tuned to the 450-kc signal.

A 450-kc signal would likewise result if the oscillator operated at 550 kc instead of 1450 kc, since the difference between 1000 kc and 550 kc is also 450 kc. In practice, though, most receiver oscillator circuits are designed to function at a frequency which is higher than that of the incoming broadcast signal.

Image Frequency

We have described briefly how the desired incoming broadcast signal is changed in frequency so that it may be amplified by an i-f amplifier. It is also possible for undesired signals to be converted to the intermediate frequency and therefore to be amplified by the i-f stages, thus causing interference. For instance, if the receiver were tuned to a weak 10,000-kc incoming signal, the oscillator normally would function at 10,450 kc to produce the desired 450-kc i-f signal. But if a strong 10,900-kc signal were likewise present in the antenna circuit, it might force its way through to the mixer along with the desired 10,000-kc

signal. Since the oscillator is functioning at 10,450 kc, it will beat with not only the 10,000-kc signal but also with the 10,900-kc signal and change the frequency of each to 450 kc, because the difference between 10,900 kc and 10,450 kc is the same as that between 10,000 kc and 10,450 kc. Both the desired and the undesired signals will then be present in the i-f amplifier and interference will result.

This type of undesired response is known as *image-frequency response*. The image frequency always differs from the desired frequency by twice the intermediate frequency. Whether the image frequency is higher or lower than the desired frequency to which the receiver is tuned depends upon the relation of the oscillator frequency to that of the incoming signal. In most cases, the oscillator frequency is higher than the desired signal frequency. Then the image frequency must also be higher, if the i-f signal is to result. In some receivers, particularly on higher frequency bands, the oscillator frequency is lower than that to which the receiver is tuned. Then the image frequency also must be lower in frequency. Thus, if the receiver were tuned to 20,000 kc and the oscillator frequency is 19,550 kc, the i-f signal is 450 kc, and the image frequency becomes 19,100 kc.

Image-frequency response is particularly troublesome at very high frequencies, since the percentage frequency difference between the image frequency and the desired frequency is small, if the i.f. is of the order of 450 kc. Under such conditions the selectivity which can be attained by the r-f amplifier is not sufficient to exclude strong undesired signals when the receiver is tuned to a weak signal. The difficulty
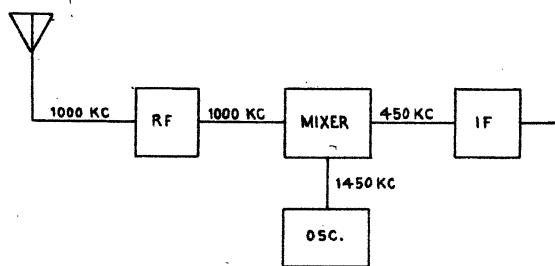

Fig. 1.—Block diagram showing how the broadcast and oscillator signals combine to form the i-f signal.

becomes more pronounced as the i.f. is lowered; that is why an i.f. of the order of 175 kc is not used in all-wave receivers. If it were, then the image frequency would differ by only 350 kc from the frequency to which the receiver is tuned, whereas if the i.f. is 450 kc, the image frequency is 900 kc removed from the desired signal frequency. Thus, in the latter case, better rejection of the undesired signal by the tuned r-f circuits is obtained.

It might seem that a still higher intermediate frequency might be employed to advantage and this is true when receivers are designed solely for u-h-f reception. Intermediate frequencies in such superheterodynes are often of the order of 3 mc or higher. For all-wave receivers, such a high intermediate frequency is undesirable because it would be necessary to omit a band of frequencies near that of the i.f. since signals in that vicinity would be picked up directly without the heterodyning oscillator. Further, lower intermediate frequencies give greater i-f selectivity and gain.

Recognition of the normal image response of superheterodynes is of particular importance in aligning all-wave receivers on short-wave bands. (See section on "Alignment" for further details.)

OSCILLATOR CIRCUITS

Examination of the radio receivers shown in Volume XV, discloses that three main types of L-C oscillator circuits are employed. These are the Hartley oscillator, the tickler-coil oscillator, and the Colpitts oscillator. Incorporated with these different oscillator circuits are many interesting and unique features that are worthy of comment. Some of these include special types of trimming and padding networks, different types of oscillator voltage injection, flattening out of oscillator bias voltage, and a host of others. Since these odd features occur in one or more of the three types of oscillator circuits mentioned, it will be most helpful to discuss the oscillator circuits themselves before analyzing the unusual features incorporated within the respective circuits.

In brief, an oscillator is a special vacuum-tube circuit wherein d-c power applied to the vacuum-tube elements is converted into a-c power, which is available from the grid or plate circuits. Stated differently, perhaps in accordance with its practical application, an oscillator is a source of alternating current and/or voltage of any desired frequency.

The ability of a vacuum tube to function as an oscillator is predicated upon its ability to amplify. Due to this behaviour, the output from a tube can be greater than its input and under proper conditions, it can supply its own input. This action contrasts the oscillator from the power amplifier. Moreover, it is this action which makes possible the development of sustained oscillations of a predetermined frequency, established by the constants of the circuit.

"Feedback" or Regeneration

The development of sustained oscillations in a vacuum tube requires the fulfillment of a number of very important conditions. The first of these is the presence of regeneration, also identified as "feedback", by which is meant the repeated transfer of a-c energy from the output circuit back to the input circuit. Any one of a number of existing paths may be provided for this feedback, there being certain preferred methods as will be shown later.

Practically, the development of sustained oscillations is instantaneous — but actually a finite amount of time is required, the time being determined by the frequency of the oscillations, the amount of feedback, and the losses in the circuit. The process of feedback is explained by the term: that some of the energy in the output circuit is fed back or returned to the input. Each time this happens, the signal in the input circuit is increased—likewise the signal in the output circuit — until eventually, which may be a matter of a very small amount of time, the a-c energy in the output circuit has reached the maximum as permitted by the operating voltages and the circuit conditions. It is important, however, to understand that the continued existence of such an oscillatory state requires the continued existence of feedback of suitable proportions. The reference to "feedback of suitable proportions" leads to another very important detail. It is possible that feedback exists in an oscillator circuit, yet sustained oscillations are not produced. Several conditions may contribute to this state.

First of all, the "phase" of the feedback is not correct. In an amplifying tube, the a-c voltage

representing the signal on the plate of the tube is 180 degrees out-of-phase with the a-c voltage which is present on the grid of the tube. In order to strengthen the signal voltage on the grid, the feedback voltage returned from the plate circuit to the grid circuit must be of the proper instantaneous polarity so as to re-enforce the signal voltage on the grid. If this condition is not fulfilled, then the feedback voltage will tend to reduce the signal voltage at the grid which in time will reduce the signal voltage on the plate circuit. The net result would be failure of the system as an oscillator. Such a system may display momentary signs of behaving as an oscillator; for that time necessary to permit the signal voltage at the grid of the tube to become zero.

Consequently, in order that the generation of oscillations start, it is necessary that the voltage fed back from the plate to the grid be in phase with the grid voltage, which means out of phase with the a-c voltage existing in the plate circuit.

Limitation of Losses

Now, it is entirely possible that properly phased feedback exists, yet the device is a very feeble and unstable oscillator. This leads to the second important condition which must be fulfilled in an oscillator system. It is the limitation of losses in the circuit and the absence of power-absorbing systems, devices, or components adjacent to or in the oscillator. The amount of feedback required in an oscillatory system to maintain oscillations depends to a great extent upon the amount of losses present in the system. By these losses are meant conditions native to the circuit which tend to conserve or dissipate the energy fed back to the grid circuit. If the coils and capacitors which comprise the grid circuit are not as efficient electrically as they should be, consequently, most of the energy fed back from the plate is dissipated in the form of heat in these components. The necessary increase in signal voltage at the grid will not be accomplished and either the system as a whole will not oscillate or it will be erratic.

Such losses are not limited to the grid circuit; they can take place in the plate circuit as well, with similar overall effects. A function of the feedback is to return sufficient a-c energy from the plate to the grid circuit which will overcome the losses present in the circuits, and also to furnish that minute amount of signal which is required to permit building up the signal voltage to a sustained oscillatory state; in other words, to provide the necessary grid excitation voltage.

Loss conditions which will prevent the development of sustained oscillations are not necessarily limited to the components which are connected directly into the oscillating circuits. It is possible that an adjacent component, such as a length of wire, or a coil with the required distributed capacity, or a tuned circuit adjacent to the oscillator circuits may be absorbing so much energy from the oscillator system at the operating frequency as to prevent its proper functioning as a source of sustained oscillations. Such energy-absorbing components or circuits adjacent to an oscillator display an effect tantamount to the introduction of very high losses into the oscillating system.

Concerning loss-producing elements in an oscillatory system, every component and every connection may be the point at which the losses reach such excessive proportions as to prevent the circuit from oscillating.

Oscillator Tubes

Defective tubes or wrong operating potentials are commonplace reasons for improper performance of oscillator circuits. As a matter of fact, changes in tube performance due to "ageing" will display a relationship to frequency of oscillation, the higher the frequency, the more critical being the requirements. In multi-band oscillator circuits, much more trouble is encountered on the high-frequency bands than on the broadcast band.

Certain tubes will generate sustained oscillations much more readily than others — even when the tubes are of identical type and manufacture. Of the various operating potentials applied to a tube, control grid, screen and plate voltage are the most important. In all tubes, the operating grid bias plays a very important role in determining the ease of starting oscillations and stability over the complete frequency range. In the case of tetrode and pentode type tubes, the screen voltage is more critical than the plate voltage with respect to stability and general output performance.

The Tank Circuit

The frequency of oscillations generated in an L-C oscillator system is influenced in part by the vacuum tube used, in that certain tubes are not capable of providing oscillations higher than a certain limit. The main factor, however, is the L-C constant of the inductance and capacity present in what is called the tuned or "tank circuit". Every oscillator which uses inductance and capacity has such a tank circuit.

The oscillator "tank" is a parallel combination of inductance and capacity, as shown in Fig. 2. It may be located in the plate circuit, grid circuit, cathode circuit or even external to the tube circuit elements as an external link which may couple the grid and plate circuits of the vacuum tube. The designation "tank circuit" is applied because of its action as a reservoir or storage tank of energy during certain moments of oscillation production.
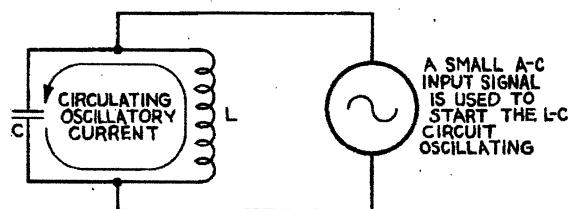
FIG. 2.—A small a-c input signal is used to start the L-C circuit oscillating.

The constants of such tanks determine the frequency of the oscillations produced in the vacuum-tube system, the actual frequency being determined by the equation

$$f = \frac{1}{2\pi\sqrt{LC}}$$

where f = the frequency of oscillations
   L = the inductance
   C = the capacitance
   π = a numerical quantity equal to 3.14
If L is expressed in henrys and
   C is expressed in farads then
   f will be in cycles per second
Also if L is expressed in microhenrys and
   C is expressed in microfarads then
   f will be expressed in megacycles.

For example, if the inductance of the oscillator circuit is equal to 250 microhenrys and the capacitance is equal to 0.00016 microfarad (or 160 micromicrofarads) then the frequency of oscillations will be:

$$f = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2 \times 3.14 \sqrt{250 \times .00016}}$$

$$= \frac{1}{6.28\sqrt{.04}}$$

$$f = \frac{1}{6.28 \times .2} = \frac{1}{1.256}$$

$$= .08 \text{ mc or } 800 \text{ kc (approximately)}$$

The inductance L may consist of one or more coils, usually one; the capacity C may consist of one or more capacitors, usually several in the oscillator system of a receiver, as will be illustrated later.

Analysis of the Tank Circuit

The analysis of how the tank circuit behaves is a very important factor in the study of oscillator circuits, and while it is true that the constants of receiver oscillator circuits are predetermined during their design and not subject to change during repair, the repairman has everything to gain and nothing to lose by understanding what happens in these circuits.

For the sake of clarity, the theoretical tank circuit will be assumed to be free of all resistance in the coil L and in the capacitor C. In other words, both these components are free of all losses which would tend to dissipate any energy fed into the circuit. If we assume that such a circuit is excited by some external force of some kind, as an a-c signal, a current of the frequency determined by the constants of the circuit will commence flowing in the tank because of a to-and-fro transfer of energy from the capacitor to the inductance. In other words, the circuit will start oscillating and theoretically would continue to do so forever, since there is no resistance or electrical friction to dissipate the electrical energy within the circuit and so damp the oscillations. Such a condition is theoretical, of course; in practice, it is impossible to achieve such a circuit completely free of resistance. The start of oscillations is brought about as follows:

Suppose we assume a momentary application of voltage to the capacitor so that it becomes fully charged; then the exciting voltage is removed. When we say a capacitor is charged,

we mean that the electrical energy stored in it is stored in the electrostatic field which exists between the active surfaces of the plates of the capacitor. This is our starting point.

1. With the fully charged capacitor connected across the coil, the capacitor begins to discharge through the coil and in so doing causes a changing current to flow through the coil. This changing current creates a magnetic field around the coil. Eventually the voltage across the coil equals the voltage across the capacitor, which coincides with the maximum current flow through the coil in, let us say, the positive direction. The electrical energy previously stored in the electrostatic field between the capacitor plates now is stored in the electromagnetic field surrounding the turns of the coil.

2. Since an electrical pressure no longer exists across the capacitor to cause continued flow of current through the coil, and since the electromagnetic lines of force have reached their maximum strength across the coil and there is no electrical force present to keep them there, the field starts to collapse. This collapsing field induces an E.M.F. across the coil in such a direction as to charge the capacitor in the opposite direction. When the capacitor is fully charged in this opposite direction, all the energy stored in the electromagnetic field around the coil has been returned to the capacitor and is in its electrostatic field.

3. The capacitor now is fully charged in the opposite direction to what it was at first, and it commences to discharge, causing current to flow through the coil in the opposite direction and the electromagnetic field around the coil begins to build up. The electromagnetic field continues increasing until it reaches its maximum strength which corresponds to the discharge of the capacitor, at which time all the energy previously stored in the electrostatic field of the capacitor now has been stored in the electromagnetic field of the coil.

4. The field around the coil starts to collapse and induces a voltage which begins to charge the capacitor in the opposite direction, eventually charging it to the full voltage and transferring all the energy in the electromagnetic field around the coil to the electrostatic field created between the capacitor plates.

Thus a cycle of current flow between the capacitor and the coil is completed, which is the same as saying that a cycle of oscillation has been completed. In a circuit having no resistance, an initial charge applied to the capacitor would cause the interchange of energy from coil to capacitor for an indefinite period. Since all circuits possess resistance, however, and this would tend to dissipate the energy, periodic excitation of such a circuit is necessary to maintain the oscillations at a constant amplitude, otherwise they would gradually decay and eventually be non-existent. The reference "tank" applied to such a circuit is due to the fact that it acts as a reservoir of energy in between the moments when the excitation is applied.

The number of times each second that the transfer of energy takes place between the coil and the capacitor, is determined by the constants of L and C. By suitable selection of L and C, the tank can be made resonant to any desired frequency and thus produce oscillations of that frequency. Oscillations may be produced in a resonant circuit, even when the excitation may be at some other frequency; all that is necessary is that the circuit be excited periodically so as to maintain the flow of current.

The Tickler-Coil Oscillator

One of the most common types of oscillators used today is the tickler-coil oscillator circuit, a typical circuit being shown in Fig. 3. The operation of the circuit is very simple and is as follows:

Initially the grid of the tube is not biased and it is considered at zero potential. In other words, the smallest voltage appearing on the
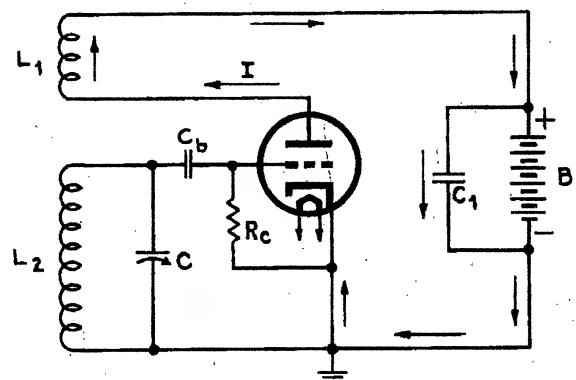

FIG. 3.—Typical tickler-coil oscillator circuit.

grid causes an immediate change in the plate current and, therefore, an a-c signal appears in the plate circuit of the tube. This a-c signal current will have to flow through coil L_1 and then back to the cathode circuit to complete its path. Coil L_1 is situated near coil L_2 in such a manner that there exists a mutual inductance between the two coils. All that coil L_1 needs is a very small quantity of a-c signal to flow through it in order to set up changing magnetic lines of flux. In other words, the most minute current change through L_1 will induce, by means of regular transformer action, a voltage into coil L_2. This small amount of induced voltage immediately excites the L_2-C network into oscillations at its natural frequency. These oscillations are then transferred to the grid of the circuit through C_b and they become the a-c signal that constantly appears on the grid. The tube amplifies this a-c signal and it always reappears on the grid with its frequency determined by the resonance of the L-C network. By virtue of the tickler-coil L_1 being in the plate circuit, some energy is always fed back from the plate circuit to the grid circuit by means of induction. This fed-back energy is supposed to replenish the energy lost because of the inherent resistance in the circuit and hence oscillations are always maintained. The resistance R_c serves as the grid resistor of the circuit and the d-c drop across this resistance is the self bias for the oscillator tube. In fact, one quick way to test if the circuit is oscillating is to find if there is any d-c drop across this grid-bias resistor. During oscillation, a d-c voltage will exist across the grid leak. This voltage is frequently mentioned as reference service data. This measurement must be made with a vacuum-tube voltmeter. The frequency of oscillations in this type of circuit is determined primarily by the inductance of L_2 and the capacitance of C. The capacitor C_1 is used to bypass the a-c signal currents around the B supply so that there will not be any voltage drop due to the impedance of the B supply.

The tickler-coil oscillator shown here has the tuned L_2-C circuit in the grid side and the tickler-coil in the plate side. There are, however, other tickler-coil oscillator circuits that have the tuned circuit in the plate side of the tube. In order to differentiate between the two, the latter type circuit is known more fully as the "tuned-plate tickler-coil oscillator" and

the circuit of Fig. 3 is known as the "tuned-grid tickler-coil oscillator."

The Hartley Oscillator

Another very common type of oscillator circuit is the well-known Hartley oscillator. There are two different types of Hartley oscillator circuits: the series-fed Hartley and the shunt-fed Hartley. As far as this section is concerned, we will consider only the series-fed circuit since that is the type most commonly found in radio receivers of today. A typical (series-fed) Hartley oscillator circuit is shown in Fig. 4. The circuit essentially consists of a
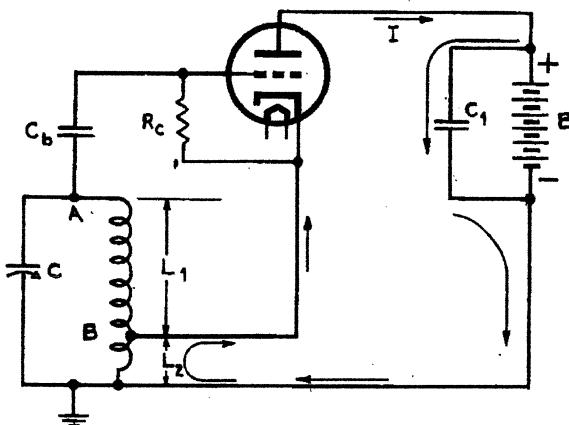
Fig. 4.—Illustrating a typical Hartley oscillator circuit.

parallel-tuned inductive-capacitive network with the coil tapped. The tapped section of the coil is returned to the cathode of the tube, the high side of the network is connected to the grid circuit, and the low side of the network is usually connected to ground or B minus. This circuit functions in the following manner.

First, an a-c signal appears on the plate of the tube in a manner similar to that described in the preceding section on the tickler-coil oscillator. The lower portion of the coil L_2, serves the same purpose as the tickler coil of the former oscillator; therefore, the inductance of L_2 is small and the tap on the coil is made very near the grounded side. The a-c signal current flowing through the plate circuit of the tube flows through L_2 to complete its path back to the cathode. The capacitor C_1 bypasses these a-c signal currents around the B supply and thus prevents a drop in a-c signal voltage due to impedance of the B supply.

The signal current flowing through L_2 sets up a magnetic field whose lines of flux envelope the coil L_1 and, therefore, a voltage is induced in this latter coil. This small amount of a-c signal induced in the coil L_1 starts oscillations of the inductance-capacitance network which consists primarily of the capacitance C and the total inductance of L_1 plus L_2. The reason why both L_1 and L_2 are involved is due to the fact that one end of the capacitor C is grounded as well as L_2. The oscillations are then fed into the grid circuit through the capacitor C_b. The grid circuit then varies according to the oscillations produced and there is a continuous replenishing of energy back to the grid circuit from the plate circuit (by virtue of the coil which is connected as an autotransformer) to sustain oscillations. In the circuit of Fig. 4, the B battery is in series with the plate coil L_2 (that coil which feeds back the a-c signal from the plate to the grid circuit), therefore, this circuit is termed a "series-fed Hartley oscillator." The resistor R_c serves the purpose of a grid-leak resistor and the tube receives its self bias due to the d-c voltage drop across this resistor. Noting whether the required amount of d-c voltage drop is across R_c, is a quick way of determining if the circuit is oscillating.

The greater the number of turns in the feedback coil section L_2 then the greater will be the voltage fed back to the coil L_1. This would mean that for a greater feedback voltage the whole coil would have to approach being center-tapped rather than tapped just a few turns from the grounded end. Although a larger feedback voltage is often desired, it is not preferable to raise the tap on the total coil to obtain this increase. This is because if the tap were placed higher up, then the capacitive reactance of C will be effectively smaller than the inductive reactance of the feedback coil L_2 and the capacitor will then offer a lower impedance to the feedback voltage. In this case, the voltage that would be fed back would be practically shunted by the capacitor C (not much going through L_2) and it would be of a degenerative nature and buck, not aid, the a-c signal output from the tube.

The Colpitts Oscillator

In the Hartley circuit, it was shown how the ratio of the two inductances of the coil was a

determining factor in maintaining oscillations. A similar situation exists in the Colpitts oscillator, of which a typical circuit is shown in Fig. 5. In this circuit instead of a tapped coil, we effectively have a tapped capacitor. The operation of this circuit is similar to the Hartley. The tube also needs the slightest amount of electron movement to change the zero bias existing on the grid and therefore to have an a-c signal appear on the plate. An a-c signal current flows from the plate of this tube through the circuit as shown and returns to the cathode to complete its path. C_1 is the feedback capacitor used to feed the output voltage back to the frequency determining components, which in this case are the coil L and the series capacitance combination of C in series with C_1. The reactance of C_1 is small compared to the reactance of the coil L so that the signal fed back is not shunted by the coil. The capacitor C_1 is of such a value that all it needs is a small amount of signal current flowing through it to start the circuit resonating. The ratio between C and C_1 is so chosen that the energy fed back from the plate through the capacitor C_1 is of such a value that there is built up enough of an oscillator voltage to be in-
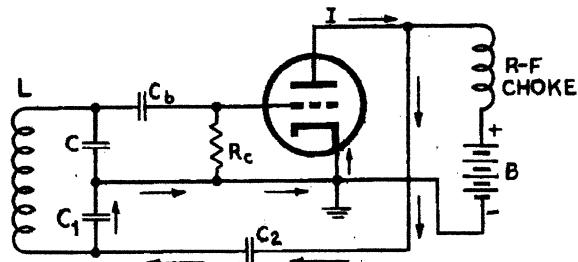
Fig. 5.—A typical Colpitts oscillator circuit.

jected into the grid to keep the complete circuit in oscillation. The r-f choke in the circuit prevents the a-c signal currents from flowing through the B supply and at the same time capacitor C_2 serves the purpose of coupling the a-c signal from the plate to the feedback capacitor C_1. The resistor R_c serves the purpose of a self biasing grid-leak resistor for this oscillator.

Galvin Motorola Model 405

In the Motorola Model 405, a Hartley oscillator circuit is used but the necessary inductance for the oscillator frequency is ob-

tained in an unusual manner. The oscillator circuit, shown in Fig. 6, uses the oscillator section of a 6SA7-GT converter tube. It will be seen that there are *three* variable components comprising the oscillator circuit, these being *C10*, *L3*, and *L4*. In the usual Hartley oscillator, the inductance for the tuned tank circuit is obtained from a single coil, but in this circuit the inductance used to determine the oscillator frequency is obtained by the parallel combination of *L3* and *L4*. The selecting of stations by this set is accomplished by permeability ganged tuning. In other words, *L3* in conjunction with the tuning inductances of the r-f stages of this receiver is used to tune in the stations similar to the previous ganged-capacitor tuning.
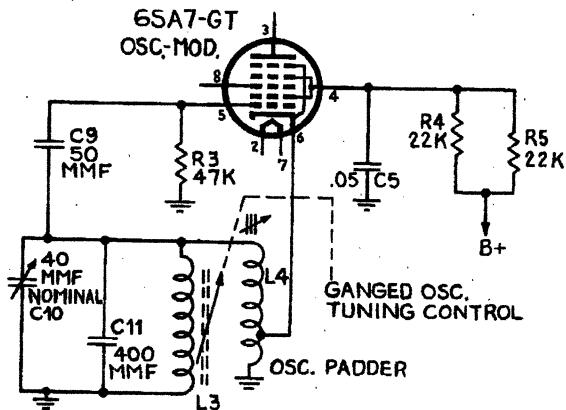
Fig. 6.—Hartley oscillator circuit of the Galvin Motorola Model 405.

Usually in such permeability-tuned Hartley oscillator circuits, the one oscillator coil that is used is tapped, the tap being connected to the cathode. In the circuit under discussion, this is not so. In parallel with the tuning inductance *L3* is another inductance *L4*, which is also made variable by a tuning slug inside its core. This latter inductance *L4* is the coil that is tapped and this tap goes to the cathode of the 6SA7-GT tube to complete the Hartley oscillator circuit. The inductance *L4* is made variable for tracking purposes. In this set it is termed the "oscillator padder" and it takes the place of the usual series padding capacitor. The slug of this coil *L4* is used for proper tracking at the low-frequency end of the dial. The total inductance necessary to determine the frequency of oscillations, as mentioned before, is obtained by *L3* and *L4* in parallel. The total inductance of any two inductances in parallel

is determined similarly to resistances in parallel, therefore:

$$L_{total} = \frac{L3 \times L4}{L3 + L4}$$

Assuming then, that if *L3* equals 300 microhenrys and *L4* equals 150 microhenrys, the total inductance would be

$$L_{total} = \frac{150 \times 300}{150 + 300}$$

$$= \frac{45000}{450} = 100 \text{ microhenrys}$$

One of the advantages derived from this type of paralleling inductances is that a more constant form of oscillator output is produced. In other words, by connecting the cathode of the 6SA7-GT tube to the oscillator padder coil *L4* instead of the oscillator tuning coil *L3*, variations in the output of the oscillator over the complete range of the band are reduced. If the hook up was such that the cathode was tapped to *L3* instead of *L4*, the mutual inductance of the oscillator coil would vary a great deal during tuning and naturally, this would disturb the frequency of oscillations. The capacity for determining the frequency of oscillations is determined by the parallel combination of *C10* and *C11*. Capacitor *C10* is used as a trimmer for the proper tracking at the high end of the band and capacitor *C11* is a 400-mmf fixed mica capacitor, this being the primary capacity-determining element of the circuit. To find the total capacity of two *capacitors in parallel*, the capacities are added similar to the way resistances in series are added. Therefore, for this oscillator under discussion the total capacity is:

$$C_{total} = C10 + C11$$

and if *C10* is trimmed to 30 mmf, then with *C11* equal to 400 mmf, we have:

$$C_{total} = 30 + 400 = 430 \text{ mmf.}$$

Knowing the total capacity in the circuit and also the total effective inductance, the frequency of oscillation is thus determined by the following:

$$f = \frac{1}{2\pi \sqrt{L_T C_T}}$$

where L_T equals the total inductance and C_T equals the total capacity in the oscillator circuit. For a more concise breakdown we get:

$$f = \frac{1}{2\pi \sqrt{\left(\dfrac{L3 \times L4}{L3 + L4}\right)\left(C10 + C11\right)}}$$

One must bear in mind, however, that the values of $L3$, $L4$ and $C10$ are variable and that after proper alignment they are considered to be fixed for the correct adjustment of the desired oscillator frequency.

Stewart Warner Model 72CR16 (9004B)

In the Stewart Warner Model 72CR16 and 72CR26 a Hartley oscillator circuit is also used with an inductive padder. The oscillator circuit is shown in Fig. 7. In this circuit the tuning of the oscillator circuit is accomplished by capacitor tuning which is ganged to the r-f tuning capacitors of the set. Only one coil is used
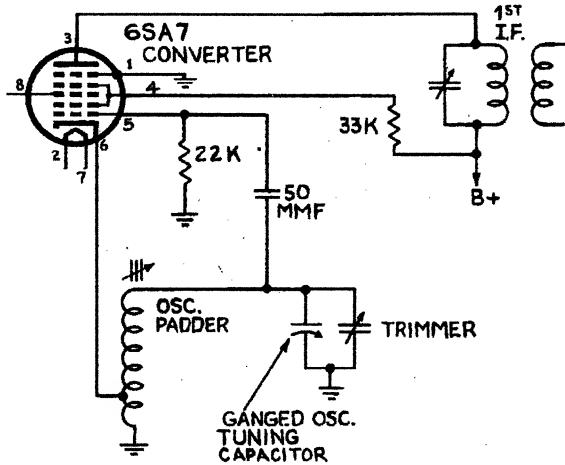
FIG. 7.—Hartley oscillator circuit of Stewart Warner Models 72CR16 and 72CR36.

in this Hartley oscillator circuit but the inductance of the coil is made variable for better tracking purposes. It is used for tracking at the 600-kc low-frequency end of the band. By the use of such permeability tuned oscillator coils, a series padding capacitor is not needed. In other words, a *series* capacitance or a *shunt* inductance in a tuned oscillator tank circuit affects the frequency of the circuit in the same way; that is, when the inductance or capacitance is increased or decreased, then the frequency of oscillations is decreased or increased respectively in either case.

This is evident by considering the simple tank circuit of Fig. 8, which shows a tuning capacitor in *series* with a capacity padder

(variable) and in shunt with a (variable) inductive padder. First, consider moving an iron slug into the core of coil L; the inductance will increase. If the inductance increases,

then from $f = \dfrac{1}{2\pi \sqrt{LC}}$, the resonant fre-

quency of the tank circuit will decrease, assuming that both capacitors remain fixed. Now if
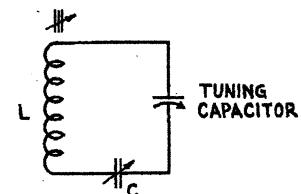
FIG. 8.—Circuit illustrating how a series capacity padder and a shunt inductance padder affect the tank circuit the same way.

both the inductance and the tuning capacitor are considered fixed and the capacitance of C increases, then the total capacitance of the circuit increases because when either one of two capacitors in series is increased, the total capacitance is likewise increased. Therefore, from the same resonant frequency formula, increasing the capacity of the circuit decreases its resonant frequency. The reader should likewise understand that capacitors in parallel and inductances in series also have their total capacity and inductance increased when any one of their components is increased. The main difference, however, between this latter type of increasing capacity and inductance compared to that illustrated in Fig. 8, is that in using the parallel-inductance method or series-capacity method, the actual increase is *small* as compared to the true change of inductance or capacitance in the unit varied.

Stewart Warner Model 62T16

In the short-wave band of the Stewart Warner Models 62T16 and 62TC36, a modified Hartley oscillator circuit is used. The circuit under discussion is shown in Fig. 9. In this Stewart Warner model, the tuning on the broadcast band is accomplished by ganged-permeability tuning, including all r-f and oscillator tuning; however, in the short-wave band there is *only* oscillator tuning but no r-f tuning. An examination of Fig. 9 shows that only the oscillator section of the converter

stage is tuned and not the r-f section. The frequency range of the short-wave band is small enough to allow the coil L in conjunction with capacitors C_1 and C_2 to be broad enough to pass *all* the frequencies of the short-wave band. Consequently, only the oscillator section need be tuned for the proper frequency conversion to take place within the 6SA7 tube.

The oscillator circuit in this band is a modified form of the Hartley circuit shown at the beginning of this oscillator section. In this circuit there are two coils in parallel to obtain the necessary inductance. One coil is used for the tuning of the oscillator and the other coil is used as the fundamental coil supplying the feedback energy from the plate circuit to the grid circuit of the oscillator section of the 6SA7 tube. The cathode in this case is grounded as well as the tapped section of coil L_1. The circuit functions in this manner:

The second grid of the 6SA7 tube functions as the plate of the oscillator tube and it is brought out of the tube, connected to another grid, to pin No. 4. Due to a momentary change of potential on the oscillator grid, pin No. 5, an alternating current flows in the oscillator section of the tube and this a.c. goes to the oscillator plate (often referred to as the "oscillator anode") of the tube. From this plate the a.c. flows through capacitor C, through the lower section of coil L_1 then to ground. Since the cathode is also at ground potential, then
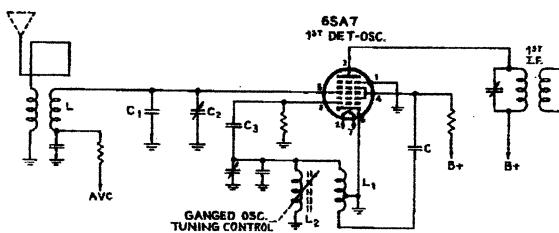
converter circuit showing tuning of the modified Hartley oscillator but without any tuning of the r-f section.

the return circuit for this a.c. is complete. By induction, a voltage is induced in the upper half of coil L_1 that starts the inductive-capacitive tank circuit into oscillation. The grid receives a voltage of oscillator frequency through capacitor C_3 and this oscillator voltage is amplified by the oscillator section of the tube. Some of this amplified signal is fed back to the coil L_1 from the anode (plate) of the oscillator

and in this manner the oscillator tank circuit is fed constantly with some feedback voltage and continues to oscillate. In this circuit the true inductance used to determine the frequency of oscillations is the inductance of the total coil L_2 in parallel with only the *upper section* of coil L_1. The reason why the inductance of the upper section of L_1 is only considered and not the inductance of the total coil is due to the fact that the tap of the coil is grounded and so are the capacitors and L_2 of the tank circuit; therefore, the tank circuit is completed through ground.

Lear Model 6614

In the Lear Models 6614, 6615, 6616, and 6619, a tickler-coil oscillator circuit is employed
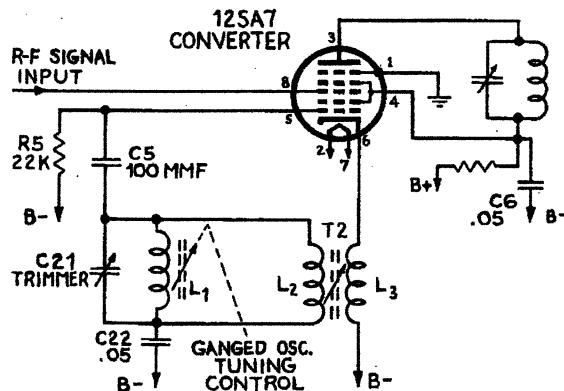
Fig. 10.—Lear Model 6614 showing tickler-coil oscillator circuit.

which uses permeability tuning (ganged to the r-f section). The oscillatory circuit is shown in Fig. 10. This tickler-coil oscillator circuit is slightly different from that shown in Fig. 3. The primary difference is that in the circuit under discussion, the tickler-coil is placed in the cathode circuit, instead of the plate circuit, of the oscillator. The a-c signal appearing on the anode of the oscillator section of the 12SA7 tube is bypassed around the B supply through $C6$ to B minus and then completes the circuit by returning to the cathode of the tube through the B minus circuit and through the tickler-coil L_3. The method of feeding back the a-c voltage from the plate circuit to the tuned grid circuit, therefore, is evident readily. In this tickler-coil circuit two inductances, L_1 and L_2, in parallel are used to obtain the necessary inductance for the tuned circuit. The inductance L_1 is permeability tuned by a movable slug

and this slug tuning is ganged to the r-f tuning section of the set, being used for station selection. The other coil L_2 and the tickler-coil L_3 are wound on a single coil-form and they are transformer coupled to each other. In other words, L_2 and L_3 represent the windings of a transformer, which is designated as $T2$ and termed the "oscillator transformer assembly." This transformer $T2$ has a tuning slug inserted into its core and its two inductances are made variable by a screwdriver adjustment. The primary purpose of the $T2$ slug tuning is to vary the inductance of L_2 and it is used as a padder for tracking at the 600-kc low-frequency end of the band. Capacitor $C21$ is a trimmer for tracking at the 1620-kc high-frequency end of the band. The frequency of oscillations for any particular fixed settings of $C21$, L_1 and L_2 is:

$$f = \frac{1}{2\pi \sqrt{\left(\dfrac{L_1\, L_3}{L_1 + L_3}\right)C21}}$$

Crosley Model 46FA

In many of the oscillator circuits used in today's radio receivers, usually it has been found that in the short-wave bands, the oscillator *bias* voltage varies over the complete tuning range of the band. This is undesired because the output oscillator voltage does not remain constant and the total oscillator frequency-response characteristic is likewise not constant. This undesirable feature is corrected by inserting a special resistance and capacitance combination into the anode oscillator circuit. This is evident when the oscillator circuit, shown in Fig. 11, of the Crosley Models 46FA and 46FB, is examined. The resistance-capacitance combination that is inserted in this circuit to overcome the above difficulty is resistor $R26$ (33-ohms) and capacitor $C15$ (0.0033-mf). Before analyzing how this difficulty is overcome, it should be understood how this undesired feature manifests itself. At first let us examine the circuit of Fig. 11, omitting for the moment the resistance-capacitance combination.

The type of circuit employed is the conventional tickler-coil oscillator with pin 6 of the 1A7GT/G tube serving as the oscillator anode and the coil L serving as the tickler-coil which supplies the feedback voltage for the tunable

oscillatory tank circuit. The a-c signal appearing at the oscillator anode is impressed across the circuit between this anode and ground to complete its path. In the usual oscillator-tickler circuit, without the resistance-capacitance network, a high-valued capacitor invariably is connected from the oscillator anode circuit to ground to complete the a-c signal return path to ground. As far as the impedance offered to this a-c signal is concerned, the inductance of the tickler-coil offers most of the impedance and practically all of the a-c signal is impressed across the tickler-coil. If the total effective impedance offered by the tickler-coil L (including any reflected impedance from the oscillatory tank circuit) would remain constant with change in frequency, then it would be satisfactory because a more constant form of oscillator voltage would be maintained. However, this is not the case. It happens that even though the inductance of coil L is very small (as are most tickler coils) the impedance it offers at the high and low ends of the short-wave bands differs appreciably enough to affect the amount of feedback voltage into the tank circuit. The change in impedance of a tickler coil used in the broadcast band is not very great, since the difference in frequencies in this band is much smaller than in the short-wave bands. For the sake of argument and to show how the impedance of L does vary in the short-wave band, let us refer to the actual frequencies used in the short-wave band in the receiver under discussion. In this short-wave band, the frequency range is 5.8 mc to 15 mc. (This is a difference of 9.2 mc compared to a difference
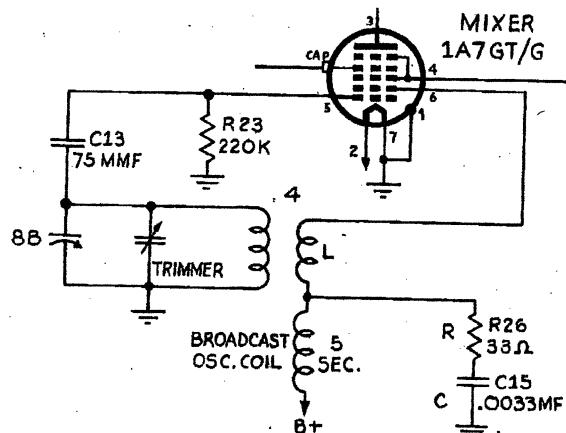
Fig. 11.—Crosley Models 46FA oscillator circuit on short-wave band showing method of obtaining flat oscillator bias voltage.

of 1.06 mc between 540 kc to 1600 kc of the broadcast band). From the equation for the inductive reactance X_L, it is found that:

$$X_L = 2\pi\, fL \text{ in ohms}$$

where f is the frequency in cycles or megacycles
  L is the inductance in henrys or
    microhenrys and
  π is a numerical value equal to 3.14

If we assume that the inductance of the tickler coil to be about 10 microhenrys, then at 5.8-mc low-frequency end of the short-wave band we find that:

$$X_L = 2 \times 3.14 \times 5.8 \times 10$$
$$X_L = 364 \text{ ohms}$$

And at the 15-mc high-frequency end of the band the inductive reactance of the tickler coil would be:

$$X_L = 2\pi\, fL$$
$$X_L = 2 \times 3.14 \times 15 \times 10$$
$$X_L = 942 \text{ ohms}$$

It has, therefore, been found that due to such relations in impedance offered by the tickler coil within the short-wave band (an impedance difference of 578 ohms) a different amount of fed-back voltage goes into the oscillator tank circuit, which in turn causes a variable bias throughout the entire band. In many instances the oscillator bias has been found to vary from —2 volts to —10 volts, due to this varying impedance of the tickler coil. It has been found, however, that certain resistance-capacitance networks compensate for this variation.

In Fig. 11, the R-C combination is just one of the ways the impedance of the anode oscillator circuit is made to vary in such a manner that at practically the whole range of the short-wave band, the voltage fed back into the tank circuit remains as constant as possible and the bias on the oscillator grid likewise remains as constant. It will be noticed that L, R and C are all in series between the oscillator anode and ground, offering a certain amount of impedance to this anode circuit. The effect of the .0033-mf capacitance $C15$ in the circuit is to offer a low impedance to the high frequencies and a higher impedance to the low frequencies. In other words, for the same frequency the impedance of this capacitor varies inversely to that of the inductance L and in that manner the total anode circuit has an effective impedance that remains more constant with change in frequency. Therefore, the a-c voltage fed back also remains more constant due

to the new load on the oscillator anode. The leveling of oscillation voltage can be determined readily by testing the bias across resistor $R23$ during the whole range of the short-wave band with the R-C circuit intact and then with resistor $R26$ open at the anode coil. It will be found that with the R-C combination in the circuit, the bias voltage will vary a small amount; while with the R-C circuit open, the bias voltage will vary a great deal. In many such oscillator circuits, there is a tendency toward production of parasitic oscillations and the 33-ohm resistor $R26$ is primarily inserted in the circuit to damp out these parasitic oscillations. This is known as a "parasitic resistor". (See section on "Parasitic Resistors.")

International Detrola Models 420 Series

A typical Colpitts oscillator circuit using a 12SA7 converter tube, is employed in the International Detrola Models 420 series, shown in Fig. 12. The primary difference between this
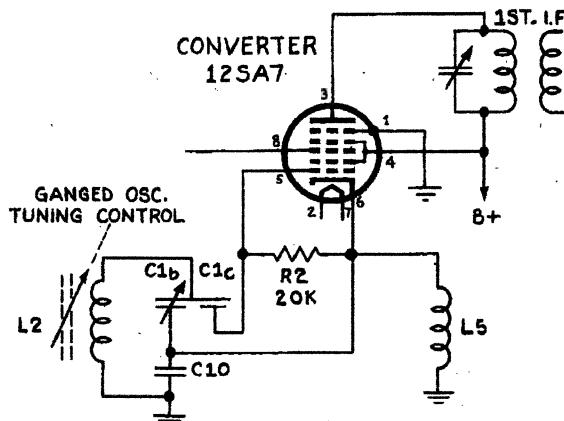
Fig. 12.—A typical Colpitts oscillator circuit found in Detrola Models 420 series.

Colpitts circuit and that shown in Fig. 5, is that the r-f choke in the Detrola receiver is in the cathode circuit instead of in the oscillator plate circuit.

The a-c signal existing on the plate of the oscillator section of the 12SA7 (the second grid from the cathode) is returned to the cathode circuit through the electrolytic filter capacitors of the B supply. The choke $L5$ prevents this returned signal from going directly to the cathode and, therefore, the signal has to pass through $C10$ to be returned to the cathode to

complete its path. Capacitor C10 serves the same purpose as the lower section of a tapped Hartley oscillator coil in that it is the feedback element in supplying the necessary a-c voltage to keep the oscillator working. Capacitors $C1_b$ and $C1_c$ are part of a dual section in which one plate is common to both capacitors. Capacitor $C1_b$ is a trimmer for aligning the oscillator section of the receiver, while capacitor $C1_c$, a fixed capacitance, is used to couple the oscillator voltage from the tank circuit to the oscillator grid. Coil L2 is the oscillator inductance and it is permeability tuned and ganged with the r-f section of the set. The choke L5, besides offering a high impedance to radio frequencies, also offers a d-c return path for the cathode circuit.

The resonant frequency of the oscillatory tank circuit is determined by the fixed value of the inductance of L2 and the series capacitance of $C1_b$ and $C10$. Therefore, the oscillator frequency determined by the resonant frequency formula is:

$$f = \frac{1}{2\pi \sqrt{L2\left(\dfrac{C1_b \times C10}{C1_b + C10}\right)}}$$

where the value for the total capacity is that determined by the two capacitors in series.

Spiegel Model 433

The Colpitts oscillator circuit of the Spiegel Model 433, shown in Fig. 13, is very similar to

CONVERTER 12SA7

1ST. I.F

GANGED OSC. TUNING CONTROL

100 MMF

20K

B+

L

.0015 MF
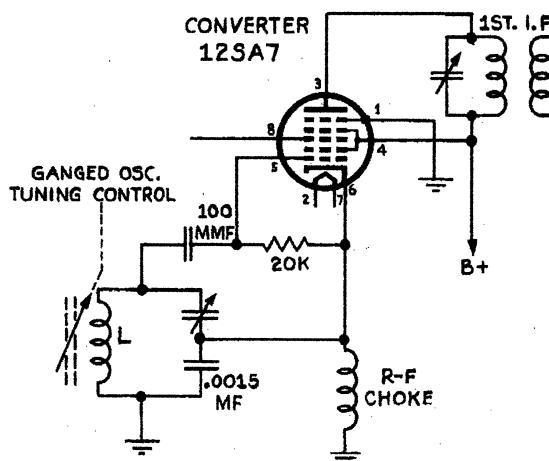
R-F CHOKE

Fig. 13.—Colpitts oscillator circuit of Spiegel Model 433.

the previous one discussed. It also has an r-f choke in the cathode circuit and uses a 12SA7 converter tube. The Colpitts circuit is evident here and consists of the feedback capacitor of .0015-mf in series with the oscillator trimmer and together both capacitors are in parallel with the ganged permeability tuning inductance L. The coupling capacitor used to couple the oscillator voltage onto the oscillator grid is a separate mica capacitor of 100-mmf. The cathode is connected to the junction between the trimmer and the .0015-mf capacitor of the tuned circuit to complete the Colpitts circuit.

RCA Model 56X

In all of the circuits discussed so far only one tube was used to accomplish the process of frequency conversion. In other words, a single tube produced the oscillations and also received the input r-f signal to mix them both within one tube envelope. There are many advantages and also some disadvantages for the use of a single frequency conversion tube. As far as we are concerned, for the moment, one of the primary reasons a single tube is used is to save space, which is always needed in radio receivers. It is a known fact, however, that when a separate tube is used to produce the oscillatory signal and another tube is used to receive the r-f signal as well as mixing both signals together, there are certain definite advantageous features offered by such an arrangement. Some important desirable features are the reduction of interaction between the oscillator and input r-f signal circuits, the ability to obtain a steady oscillator voltage, and the reduction of oscillator frequency drift.

One such type of circuit using a separate oscillator tube is in the RCA Model 56X, as shown in Fig. 14. The separate oscillator tube employed is the 12J5GT. The type of oscillator used is the usual tickler-coil oscillator with the tickler coil in the cathode circuit of the tube. Upon first examination of this oscillator circuit it appears that the tank circuit is open at one end. As far as wiring connections are concerned, one end of the tank circuit *is* open with no *direct* connection to the grid circuit of the oscillator tube. But this does not mean the grid of the 12J5GT will not receive any oscillator signal voltage. Upon closer examination of this circuit it will be noticed that

above the tank inductance *L8* is another coil *L7*, which has one end connected directly to the grid of the 12J5GT tube and the other end open. This type of wiring is purposely made because between the open end of *L7* and the high side of the oscillator coil *L8* an *effective capacity* exists which is used to couple the oscillator voltage onto the grid of the 12J5GT tube. This type of a coil, *L7*, is termed a "gimmick" and for a more complete discussion on "gimmicks" of all types see the section on "Gimmicks" in this book.
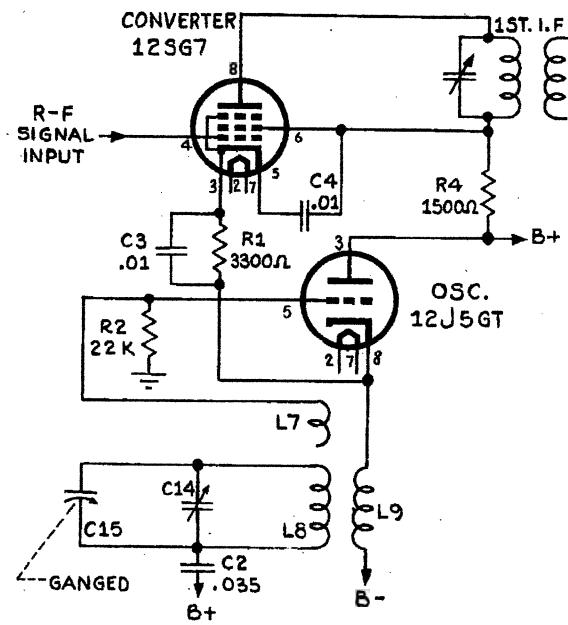
Fig. 14.—RCA Model 56X using a separate oscillator tube.

Now let us see how the oscillatory voltages are injected into the 12SG7 tube to be mixed with the input r-f signal for the necessary frequency conversion. It has been found that when using a separate tube as an oscillator and a separate pentode tube for mixing purposes, injection of the oscillatory voltage into the cathode circuit of the pentode tube gives overall characteristics desirable for good frequency response. In this respect it will be noted in Fig. 14 that the oscillator signal is injected into the cathode circuit of the 12SG7 tube through *C3-R1* combination from the cathode circuit of the 12J5GT oscillator tube. The incoming r-f signal mixes with the oscillator signal in the 12SG7 tube to produce a number of additive and subtractive frequencies of which the first i-f transformer selects the desired frequency.

Stewart Warner Model 61TR36

In the Stewart Warner Models 61TR36 and 61TR46, a separate oscillator tube is also used, but instead of a tickler-coil oscillator, a Hartley
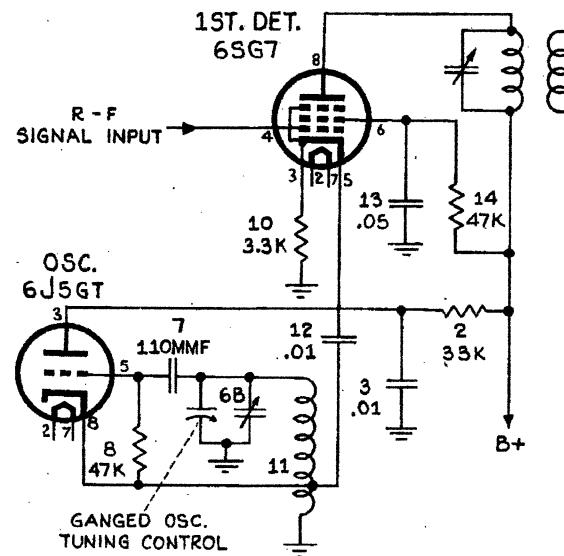
Fig. 15.—Stewart Warner Model 61TR36 using a Hartley oscillator with separate oscillator tube.

oscillator is employed. The circuit is shown in Fig. 15. A 6J5GT tube is used as the oscillator and a 6SG7 tube is used as a first detector to accomplish the frequency conversion. The oscillator uses a conventional Hartley circuit with coil *11* tapped at the low end and the tap going to the cathode of the 6J5GT tube. The 47,000-ohm resistor *8*, is used as the oscillator bias resistor while the 110-mmf capacitor *7*, is used to couple the oscillator voltage to the grid of the oscillator tube from the tank circuit. The oscillatory tank circuit consists of the ganged tuning capacitor with its attached trimmer (part *6B*) in parallel with the complete inductance of the coil (*11*).

The oscillator voltage is injected into the cathode circuit of the 6SG7 tube through the .01-mf capacitor *12*, from the tap on the oscillator coil, which is considered as the same point as the cathode of the 6J5GT tube. The same benefits of using a separate oscillator tube and a separate pentode mixing tube with cathode injection is realized here as in the preceding discussion. The 12SG7 tube used in the preceding discussion and the 6SG7 tube at first appear as though they are drawn wrong by having two cathode leads coming out of the

tubes; in fact, that is the way these tubes are manufactured: the cathode circuit is brought out of the tube at two different pins.

Stromberg-Carlson Models 1020, 1120

In many oscillator circuits, one of the biggest causes of trouble is that the frequency of oscillations tends to drift, especially when the tubes and elements of the set are not heated completely. In the Stromberg-Carlson Models 1020, 1120, a unique feature is incorporated in that a special compensating network is used to stabilize the oscillator frequency. The oscillator circuit with its compensating network *RC1* is shown in Fig. 16. The oscillator circuit employed is the usual Hartley circuit. The tap on the Hartley oscillator coil is returned to the cathode of the 6SA7 tube through part of the short-wave oscillator coil *L7*. This is done primarily to facilitate switching arrangements. The compensator *RC1* consists of an 8-mmf capacitor and a 39-ohm resistor. The capacitor is wired so that it is across the oscillator tuned circuit and, therefore, its capacity is an effective part of the oscillator tank circuit. In other words, *C12*, *C5*, *C14*, and the 8-mmf capacitor

Fig. 16.—Broadcast-band oscillator circuit of Stromberg Carlson Models 1020, 1120 showing compensator network.

in conjunction with the inductance comprise the components that determine the frequency of oscillations. The 39-ohm resistor is connected across the 6.3-volt filament circuit of the set, the voltage of which is supplied by a 6.3-

volt filament winding on the main power transformer. The part that the compensator network plays is as follows:

As the oscillator tube filament heats, all the elements relative to the oscillator section of this tube are likewise warming up. During this warm-up period a physical expansion of the oscillator tube elements occurs. Due to this expansion the oscillator frequency tends to drift away from the setting to which the set is tuned and, therefore, this oscillator drift must be stabilized in order that the set stay tuned to the same station. This is accomplished by the *RC1* compensator network. During the time required for the oscillator circuit to warm up and operate under a constant temperature, the *RC1* network compensates for the drift in frequency by gradually changing the capacity of the 8-mmf capacitor of the *RC1* network. The 39-ohm resistor, being across the 6.3-volt filament circuit, gradually heats up in accordance with the warm-up period of the 6SA7 filament as well as the filaments of all the 6.3-volt tubes. This resistor and 8-mmf capacitor are built together in a separate unit and they are situated close to each other. The gradual heating of the resistor imparts to the 8-mmf capacitor a gradual increase in temperature. This increase in temperature gradually changes the value of the capacity of the 8-mmf capacitor, which in turn changes the frequency of oscillations. What we have then during the warm-up period is the drifting in frequency of the oscillator which is then stabilized by the change in capacity of the 8-mmf capacitor due to the heating of the 39-ohm resistor. The reason the 39-ohm resistor was placed in the *filament* circuit is evident readily if one realizes that all warming up is done by the heating of the tube filaments. Consequently, the heating of the 6SA7 oscillator filament and the 39-ohm resistor is accomplished at the same time so that the drift in frequency can be immediately stabilized.
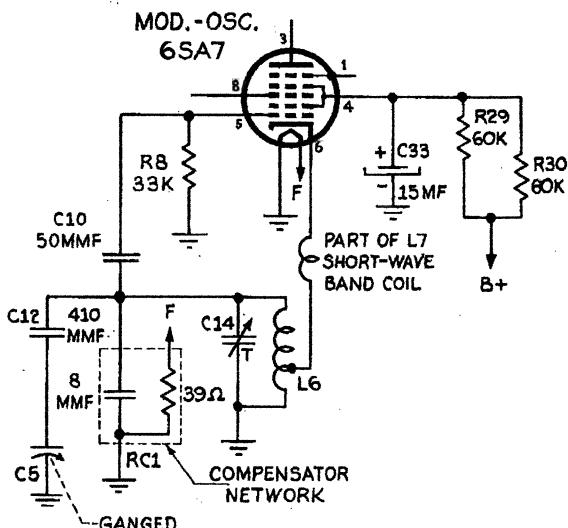
RCA Model 56X10

One of the past difficulties with radio reception on certain short-wave bands was the difficulty in getting enough bandspread. In the RCA Model 56X10, the frequency coverage of the short-wave band is 8.9 to 12.0 mc, a difference of 3.1 mc, and in order to have proper

tuning on this band, bandspreading is provided. This bandspreading is accomplished in both the r-f and the oscillator circuits. For our purposes, the oscillator bandspreading will be discussed here and its circuit is shown in Fig. 17.

The circuit uses a 12SA7 pentagrid tube employing a form of tickler coil oscillator with the tickler coil L8 in the anode circuit of the oscillator section of the 12SA7 tube. The coil L6 is the broadcast tickler-coil oscillator and serves no immediate purpose but is left in the circuit to facilitate switching arrangements. The other section of the broadcast tickler-coil oscillator is not shown because when the band switch of this set is turned from the broadcast band to the short-wave band, the .01-mf capacitor C13 effectively shorts this coil out. That is, the reactance of the .01-mf capacitor C13 that appears across this broadcast oscillator coil offers a very small impedance compared to the reactance of the circuit it shorts. In order to accomplish the necessary requirements of bandspreading on the short-wave band, the oscillator coil L9 is tapped, with the trimmer and tuning arrangements connected to this tap and the oscillator grid going directly to the high side of coil L9. From this circuit it may be seen that the total inductance of the coil L9 is not used to determine the frequency of oscillations. In fact, the tank circuit consists of capacitors C27, C15, C28, the lower half of coil L9, and those reflected impedances

into this tank due to the autotransformer coupling of the upper half of coil L9 and trimmer C29.

By this circuit providing greater bandspread, we mean that the tuning capacitor C27 has to be rotated somewhat further than usual to tune from one short-wave station to another. In the 8.9-to-12-mc short-wave band, without some means of bandspreading, the stations would be too crowded on the tuning dial. Tapping down on L9 with the tuning capacitor, reduces the L-to-C ratio of the tank circuit and according to this reduction, the tuning capacitor has to be rotated quite a bit in order to make the necessary change in frequency. This can be understood upon examining the resonant-frequency equation,

$$f = \frac{1}{2\pi\sqrt{LC}}$$ Since the

L-to-C ratio of the tank circuit is reduced, that means a larger change in C is needed to produce the necessary change in frequency. Since the tuning capacitor is primarily represented by the C in the equation, then the tuning capacitor must have a greater change in capacity (i.e., rotated more) in order to obtain the necessary change in frequency when going from one station to the other.

Capacitor C28 is a trimmer across the main tuning capacitor C27, but it is used for tracking only at the 1300-kc high-frequency end of the broadcast band. In order to obtain the necessary tracking at the high-frequency end of the short-wave band, this receiver uses a 190-260-mmf trimmer C29 across the whole oscillator coil L9. This C29 trimmer, in conjunction with the upper half of coil L9, reflects a certain impedance into the tuned circuit. One should remember that the upper and lower halves of coil L9 are autotransformer coupled to each other; consequently the reflected impedance is determined primarily by the ratio of the number of turns between both parts of coil L9 and the amount of capacity offered by C29. It should be noted that a change of capacity in C29, of say about 30-mmf, does not mean a 30-mmf change in the tank circuit capacity; such a change only means a relative change in the capacity of the tank circuit. This relative change is determined mathematically by a number of factors involved in this circuit, but it is too complex to discuss here.
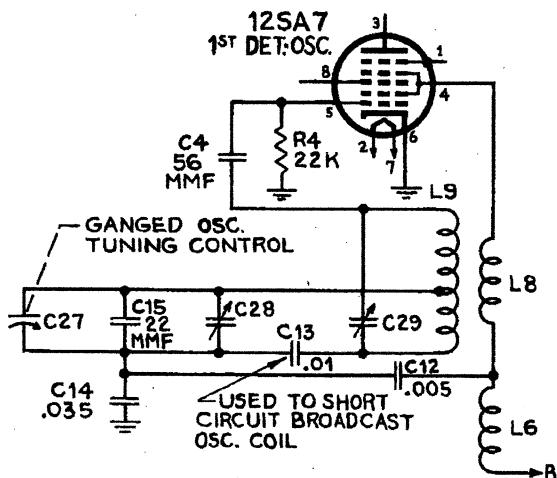
Fig. 17.—Short-wave band of the RCA Model 56X10 showing how they provided for bandspread on short wave.

Stewart Warner Model 9001C

In the Stewart Warner Models 9001C, D, E, F, use is made of three different types of oscillator circuits, i.e., the Colpitts, Hartley, and tickler-coil oscillators. The Colpitts circuit is used in their push-button tuning arrangements, the Hartley circuit in the broadcast band, and the tickler-coil circuit in the short-wave band. These three oscillator circuits are illustrated in Figs. 18A, B and C. There are quite a few interesting features about these sets that require separate consideration.

First of all if Figs. 18A and 18B are examined, it will be noted that in the broadcast band, Fig. 18A, a Hartley oscillator is used
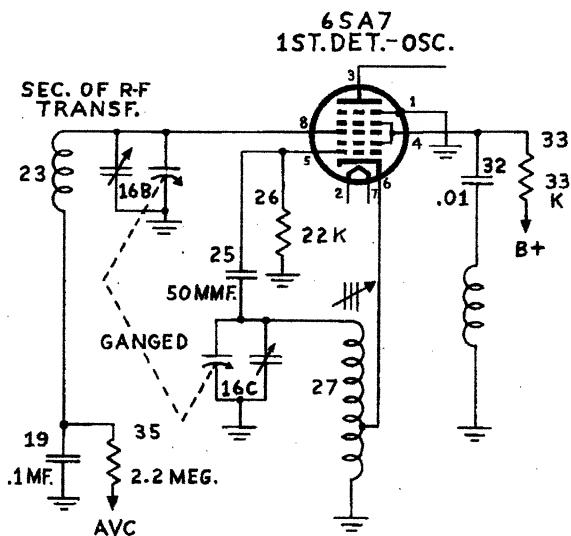
Fig. 18A.—The Hartley oscillator circuit used in the broadcast band of the Stewart Warner Models 9001C, -D, -E, -F.

whereas in the short-wave band, Fig. 18B, a tickler-coil oscillator is employed. Why this change is made is due to the type of tube used as an oscillator. When a Hartley oscillator is used with a 6SA7 tube, a fairly high impedance is in the cathode circuit of this tube. Due to this high impedance, the 6SA7 tube was found to be unstable as an oscillator on the short-wave bands, although the tube will operate quite satisfactorily on the broadcast band. Therefore, for the same 6SA7 tube two different types of oscillator circuits are used on two different bands for the primary reasons of stabilizing the oscillator frequency on the short-wave band.

Another very interesting feature in this set is the unique way one tuning capacitor is used for both r-f tuning and oscillator tuning on two different bands, as may be seen in Figs. 18A and 18B. The ganged tuning capacitor, part No. 16B, that is used for the r-f tuning of the 6SA7 tube on the broadcast band is also used as the oscillator tuning capacitor in the short-wave band. In this unique switching arrangement, the tuning capacitor 16B functions in such a manner, in conjunction with its associated switched in circuit elements, so as to tune in the desired stations on either band. It will be noticed that in the broadcast band, the capacitor 16B (together with its trimmer) is used with only the secondary of the r-f transformer 23 to form its r-f tuned circuit. In the shortwave band, however, the same tuning capacitor is used with a different inductance (namely, the secondary of transformer 31) and capacitors 28, 29, and 30, to form its complete oscillator tank circuit.

In all of the Stewart Warner models having push-button tuning, a Colpitts oscillator is employed. The Colpitts oscillator circuit used in this set is illustrated in Fig. 18C. The push button for this particular station selection, when depressed, switches capacitors 13 and 14 as well as coil 8 into the 6SA7 oscillator circuit. Most of the regular broadcast and short-wave band components used for their respective tunings are omitted by the special switching
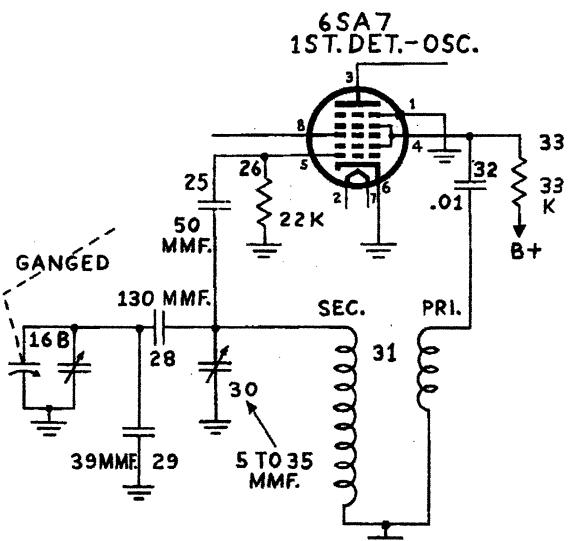
Fig. 18B.—The tickler-coil oscillator circuit used in the short-wave band of the Stewart Warner Models 9001C, -D, -E, -F.

arrangement incorporated with the push-button tuning. It will be found, nevertheless, that the lower half of the broadcast Hartley oscillator coil *27*, is still in the cathode circuit of the 6SA7 tube. This coil is left in the cathode circuit to facilitate switching arrangements and to provide a ground return path for the cathode. The Colpitts oscillator circuit is identified here with coil *8* in parallel with capacitors *13* and *14*, the junction of which goes to the cathode of the tube through ground. The functioning of this oscillator circuit is that the a-c signal appearing on the plate of the oscillator anode is bypassed around the B supply by the .01-mf capacitor *32* going to the 1000-
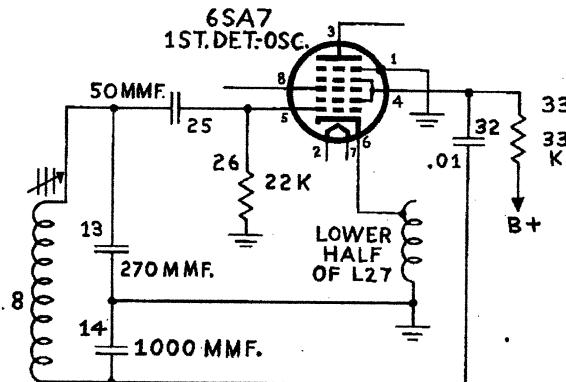
Fig. 18C.—The Colpitts oscillator circuit used in the push-button tuning (1st position) circuit of the Stewart Warner Models 9001C, -D, -E, -F.

mmf capacitor *14*. This latter capacitor *14* is the feedback capacitor that helps start the Colpitts tank circuit oscillating.

NOTES ON OSCILLATOR TESTS

Superheterodyne Oscillator Signal Voltage

For proper operation, the oscillator signal voltage applied to the mixer circuit should be far greater than the strongest incoming broadcast signal. The actual value of the oscillator voltage is dependent upon the type of mixer circuit employed. If a pentode mixer is used, and it is designed to operate at a grid bias of —10 volts, the *peak* oscillator signal voltage applied to the mixer grid or cathode circuit should not exceed —9 volts. For any other value of mixer grid bias, the peak oscillator voltage applied similarly should be 1 volt less than the grid bias.

These limitations as to the magnitude of the

oscillator signal voltage so applied arise from the fact that, if the oscillator voltage exceeds that of the grid bias, the mixer will draw grid current and load the r-f tuned circuit, resulting in reduced gain and distortion.

In many of the receivers, the tube that performs the function of frequency conversion has to have a correct amount of peak oscillator signal voltage, depending upon its operating voltages. If the peak oscillator voltage for efficient operation is about 12-volt minimum, then if this voltage becomes higher, no harm will be done but if this voltage is too low, the i-f signal output will be subnormal. In some all-wave receivers, which operate up to ultra-high-frequency bands, only a portion of the total oscillator signal voltage available normally is employed. While some conversion gain is sacrificed thereby, better overall performance is obtained.

Testing Superheterodyne Oscillators

To check an oscillator for operation, we may employ several methods. The simplest one is to measure the rectified voltage developed across the oscillator grid resistor. If the oscillator is functioning, a d-c voltage will be developed at the grid which is negative with respect to the point to which the grid returns. This voltage will vary to some degree over the band, but normally this variation will not be greater than a ratio of 2 to 1 although variations of 50% or more may be found in some receivers. If the voltage drops to zero or becomes slightly positive, it is an indication that the oscillator is not functioning. If the voltage is higher than the usual values and a pentagrid mixer is employed, satisfactory operation will still be secured.

It is also possible to check for operation by a current measurement. A 1-ma meter may be inserted in series with the oscillator grid resistor at its return point to cathode or ground. The current reading which thus results may be converted to represent the voltage reading which would be obtained if an electronic voltmeter were connected across the oscillator grid leak. This is done by Ohm's law, $E = IR$. Thus, if the oscillator grid leak is 50,000 ohms and the meter reads 0.2 ma, the average rectified voltage across the oscillator grid leak is 50,000 $\times$ 0.0002 or 10 volts.

A meter may also be used in series with the plate-supply (oscillator anode) voltage lead to the oscillator tube to indicate oscillation. When so connected, the plate current is shown. If the oscillator is functioning, touching its grid will cause an appreciable change in plate current; if not, there will be little or no change in current reading.

Uniformity of oscillation may be checked by rotating the gang capacitor over the tuning range and watching the change in current reading. If the oscillator becomes inoperative, the grid current will drop to zero. The grid current normally will vary slightly over the operating range but in normal operation the variation should be about 2 to 1.

Sudden dips in current or voltage as the oscillator frequency is varied are caused by absorption, due to the proximity of other circuits which have resonant frequencies corresponding to the frequency to which the oscillator is tuned. It is very important not to disarrange the oscillator circuit wiring during repairing or testing, since some of the oscillator energy may thus be coupled to another circuit from which it is isolated normally, resulting in dips, frequency shift and occasionally stoppage of oscillation.

While stoppage of oscillation will prevent the receiver from functioning as a superheterodyne, it is still possible that signals may be received. A strong local signal near the low-frequency end of the standard broadcast band may reach the receiver with sufficient intensity to pass through the i-f amplifier directly. This is particularly the case with small midgets, which employ no r-f stage, and in which only a single i-f stage is employed, tuned to about 465 kc. In such cases, the signal may be received at any point on the dial.

It is possible that the oscillator may be functioning but at a frequency remote from the proper point, so no signal at the intermediate frequency will result. It may also happen, if a separate oscillator is employed, that an open or short circuit in the coupling to the mixer may result in an extremely weak signal, even when the oscillator is functioning at normal frequency and output.

Troubles of these types are best checked with signal-tracing instruments, which are designed for these specific tests. However, it is possible

to check the oscillator frequency by using an auxiliary all-wave receiver, coupling the antenna lead loosely to the receiver oscillator under test, and rotating the all-wave receiver tuning capacitor over each band until the oscillator signal is picked up. A tuning indicator of some type is a desirable accessory for the all-wave receiver used for this purpose, since the oscillator signal is unmodulated, but a *thump* will be heard in the speaker if the radiated oscillator signal is strong enough.

It is also possible to inject an unmodulated signal from a test oscillator in the mixer circuit and thus restore operation. This serves as an indirect check on the oscillator frequency and its coupling to the mixer. The receiver under test should be tuned to a strong local signal. If tuned to 600-kc and the i.f. is 450 kc, the test-oscillator signal frequency should be set at 1050 kc.

Hum and Other Troubles

Other oscillator troubles may be present in which additional information may be required regarding the characteristics of the oscillator signal. It is possible that the oscillator signal may not be a pure, unmodulated wave such as is required for perfect operation. If the filtration of the voltage supply for oscillator operation is inadequate, hum may appear on the oscillator signal. This hum will then modulate any r-f signal present in the mixer, though the hum may not appear in the speaker unless the r-f signal is tuned in. Where there is reason to suspect this condition, the hum level of the d-c voltage supplied to the oscillator should be checked. If this is abnormally high, additional filtering should be introduced and its effect on the output of the receiver noted. If the modulation of hum disappears, then it can be assumed that it was caused by insufficient filtering of the plate supply. In this connection, proper by-passing of the heater is important, especially at the higher frequencies.

If the receiver blocks when tuned to a strong local station or when too high a signal is fed into it from the test oscillator, a variation in power-supply voltage may cause fluctuation of the oscillator voltage and, consequently, of the oscillator frequency. Occasionally, this trouble also results in "motor-boating." The method of test described above will show whether the

trouble is tied in with the local oscillator or elsewhere.

In some receivers, avc action on the r-f signal grid causes the frequency of the oscillator section of the converter to change. This is particularly troublesome at the high-frequency end of the tuning range. When a strong signal is being tuned in, especially on short-wave bands, the point of resonance is not the same when the receiver is tuned from a higher-frequency point on the dial as it is when the tuning is approached from a lower-frequency point. Often it is necessary to tune back and forth around the normal point of resonance before the station can be tuned in for maximum response. Then when the receiver is tuned a little beyond this point, the signal may drop out completely and retuning in the same manner may have to be repeated. This annoying condition occurs because a strong signal creates a high avc voltage, and the high avc voltage increases the bias on the converter input grid, thereby decreasing the mutual conductance of the converter tube. This changes the tube load on the oscillator tuning circuit and causes a frequency shift which is particularly large when the shunt capacity of the tuning system is a minimum, as it is at the high-frequency end of the tuning range.

The correction of this trouble requires some modification of the avc so that less control voltage is applied to the converter grid or reduction of the signal strength of the offending station by changing the antenna location or otherwise decreasing the signal pickup. Triode-hexode converters, such as the 6K8, are substantially free from this trouble due to the modifications in the design of the tube, but triode-heptodes, such as the 6J8, will be subject to this trouble unless precautions were taken in the design of the receiver to minimize such effects.

TUNING ASSEMBLIES

A rather radical departure from former practice is to be found in the tuning assemblies of some of the new radio receivers. In the past, only an occasional set was found in which tuning was accomplished by varying the inductance, rather than the capacitance of the tuned circuits. (Of course, such tuning has been used extensively for some time in i-f transformers, but in this section we are considering only r-f and oscillator tuning.) Since the publication of Volume XIV of the "Perpetual Trouble-Shooter's Manual," however, a considerable number of receivers have appeared in which tuning is accomplished by means of ganged variable inductors alone or a combination of ganged variable inductors and capacitors.

One of the principle reasons for the use of variable inductance rather than variable capacitance is saving of space. In addition, and particularly at high frequencies (as in the f-m bands), undesired coupling between circuits, which is sometimes produced by the common shaft of ganged capacitors, is avoided. Another advantage is that under certain conditions higher tuned-circuit Q may be obtained, and, therefore, higher gain.

Just as push-button tuning may be applied to radios using ganged capacitors, so it may be used with ganged inductors. An interesting example of this is found in the Galvin Motorola Models FD6 and NH6, which are discussed later.

Admiral Model 6A1

In this receiver, which covers the single tuning range, 540 kc—1630 kc, the plate circuit of the r-f amplifier is tuned by a variable inductor, while the loop antenna and oscillator are tuned by variable capacitors. This circuit is shown in Fig. 1.

Two principal benefits are derived from this choice of tuning elements. First, it eliminates stray coupling due to the impossibility of getting a perfectly grounded capacitor rotor. Because of the imperfect grounding, a certain amount of r-f current flows in the capacitor shaft, coupling the sections of the ganged capacitor together. If the tuning elements in

the input and output of the r-f stage are capacitors on a common shaft, the coupling through the shaft tends to make the stage unstable. By removing the coupling, the tendency to instability is likewise removed.
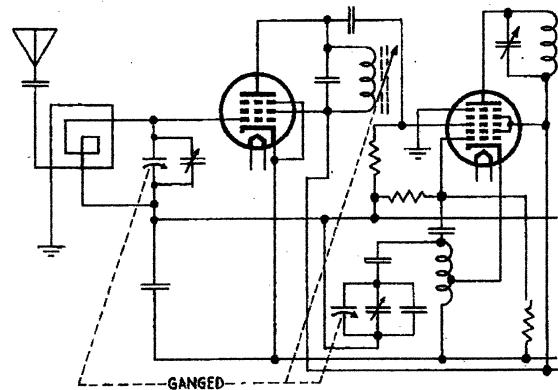
FIG. 1.—The tuning circuit of the Admiral Model 6A1 receiver, in which the loop and the oscillator are tuned by ganged capacitors which in turn are ganged with a variable inductor in the output of the r-f amplifier.

The other major advantage is found in that a tuned circuit employing variable inductance and fixed capacitance, has greater sensitivity at the low end of the band over which it is designed to tune than at the high end; whereas in a circuit using variable capacitance and fixed inductance the reverse is true. Therefore, a combination of the two provides a relatively constant response throughout the band.

Belmont Model 8A59

This set uses a total of 6 ganged variable inductors and 9 adjustable inductors, in covering the standard broadcast band and four short-wave bands. However, not all of these coils are used at once; the band switch connects only certain ones into the circuit on any one band. The tuning circuit arrangement used is shown in Fig. 2. Despite the use of a six-gang tuner, as opposed to the three-gang variable capacitor that might be used to provide tuning in a set covering the five bands covered by this one, space is not wasted, since the variable coils are much smaller than variable capacitors and at
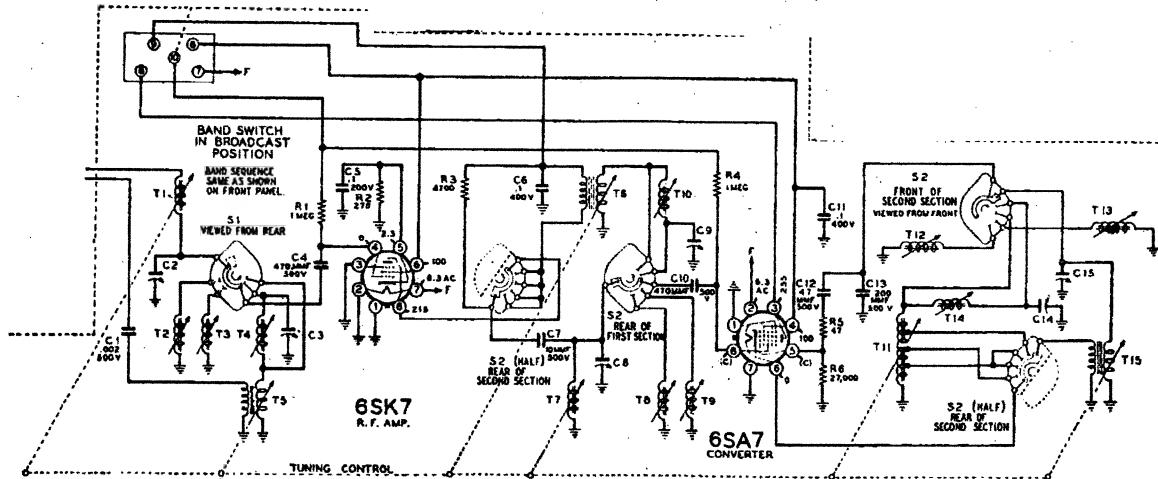
Fig. 2.—Six ganged variable inductors and nine adjustable inductors tune this five-band receiver, Belmont Model 8A59.

the same time, more flexibility in design is obtained.

The "*clarified schematics*" on Belmont pages 15-9 and 15-10, clearly show which of the ganged and adjustable coils are used in each band. In the broadcast band, *T1*, *T7*, and *T15*

are used as the antenna, r-f, and oscillator coils respectively; they are ganged and not used in any other band. The other three ganged coils (*T5*, *T6*, and *T11*) are used on the other four (short-wave) bands, but in conjunction with the nine adjustable coils.

REAR VIEW
OF CHASSIS SHOWING
TUNER UNIT & DRIVE ASSEMBLIES
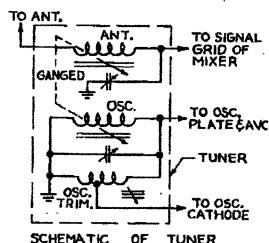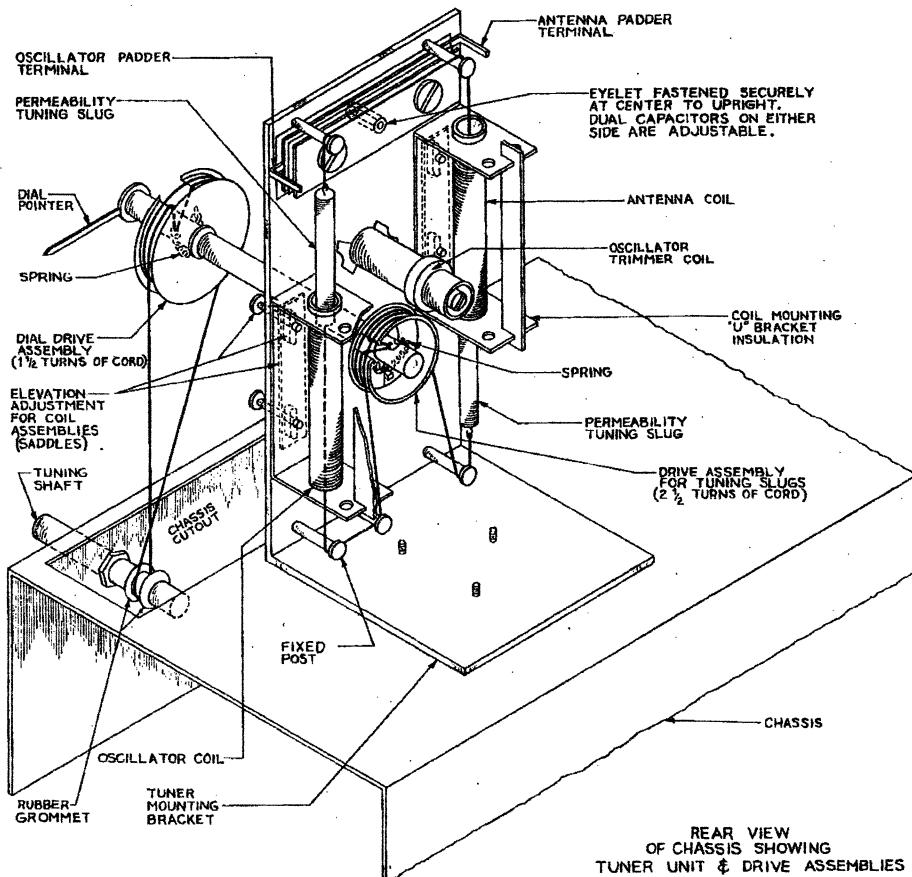
SCHEMATIC OF TUNER

Fig. 3.—Ganged inductor tuning of the "saddle" type is used in the Ray-Energy Model AD-4 receiver. The schematic diagram of the tuner is shown above.

In the 49-meter band, one adjustable coil is connected in series with each of the three ganged coils; that is, $T4$ is connected in series with the secondary of $T5$ in the antenna circuit, $T10$ is in series with the secondary of $T6$ in the r-f circuit, and $T14$ is in series with $T11$ to form the oscillator coil. In the 31-meter band, the three adjustable coils ($T4$, $T10$, and $T14$) are shorted, otherwise leaving the circuit unchanged from the 49-meter band. In the 25- and 19-meter bands, an adjustable coil is connected in parallel with each of the ganged coils instead of being connected in series as in the 49-meter band. That is, in the 25-meter band, $T2$ is connected in parallel with the secondary of $T5$, $T8$ is in parallel with the secondary of $T6$, and $T13$ is in parallel with $T11$. In the 9-meter band $T3$ is in parallel with the secondary of $T5$, $T9$ with the secondary of $T6$, and $T12$ with $T11$. The four short-wave bands are essentially the same, the only difference being the placement and use of the nine adjustable coils.

Ray-Energy Model AD-4

A type of tuner construction which will probably become quite common is found in this receiver. Fig. 3 illustrates the tuner, which is of the so-called "saddle" type. The saddles are the variable inductors inside which the movable cores are varied in position by the tuning-drive mechanism. The saddles themselves are also movable, but their positions are changed only during alignment. In addition to the saddles, the antenna and oscillator padding capacitors and the oscillator trimmer coil are mounted on the tuner. (For a description of the function of the padder coil in an oscillator tuned by a variable inductor, see the section on "Oscillator Circuits".) Despite the large number of items included in the tuning assembly, it is quite small in size. The schematic diagram of the electrical circuits of the tuner is shown also in Fig. 3.

Zenith Model 8H032

This receiver covers the standard broadcast band, the old (42 mc-50 mc) f-m band, and the new (88 mc-108 mc) f-m band. In the broadcast band, it uses conventional variable capacitor tuning, but in the f-m bands, variable inductors are used. The cores of the three-ganged inductors, which are mounted in line, are connected by long rods to a rectangular plate, as shown in Fig. 4. In addition to attaching the cores to the plate, the rods permit adjustment of the individual cores relative to each other; this adjustment is necessary in alignment.
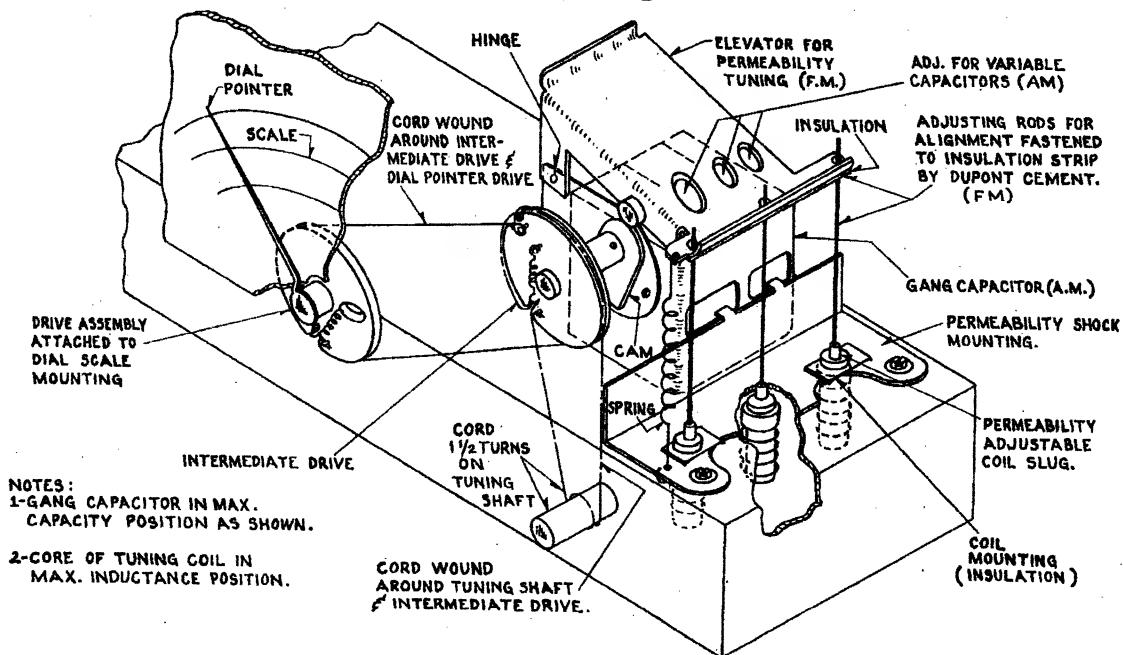
FIG. 4.—A combination of ganged variable capacitors and inductors are the tuning elements in the Zenith Model 8H032 a.m.-f.m. receiver.

One side of the plate or "elevator", which is parallel to the line of the three cores, is hinged to a bracket which is attached to the frame of the ganged capacitors. On one of the adjacent sides of the plate (that toward the front of the receiver) is a rounded projection which serves as a cam follower. The cam on which it rides is mounted on the shaft of the ganged capacitors, so that as that shaft turns in response to rotation of the tuning knob, the plate pivots on its hinge, moving the cores in or out of the tuning inductors. The plate is held against the cam by a coil spring.

Motorola Model FD6

An ingenious solenoid-operated tuner, Model ST-54, is used in the Motorola Model FD6 auto radio, and also in several other models in the 1946 Motorola line of auto radios. The tuning cores of a three-gang variable-inductor tuner, one of which is shown at A in Fig. 5, may be set automatically to any one of five pre-selected positions (push-button tuning) or positioned manually by the tuning knob (manual tuning).

Briefly, this is accomplished by means of a single-pole, six-position rotary switch, shown at B which is rotated by the solenoid until it opens the solenoid coil circuit in a position selected by one of the six pushbuttons at C (one of which is for manual tuning). At the same time that the switch is rotated, a movable plate D is carried back and forth by the solenoid plunger O. When the action of the plunger ceases, the plate is pulled by a spring E against one of six movable stops G; which one it is depends upon the position of the rotary switch. Since the cores of the tuning inductors move with the plate, the tuning of the radio is determined by the position of the stop against which the plate rests. Five of these stops are preset when the pushbuttons are set up, while the position of the sixth is controlled by the manual tuning knob.

Now let us consider the operation of this device in more detail. Fig. 5 shows the electrical circuit and some of the mechanical features of the tuning-drive mechanism. When a pushbutton at C is depressed, the power is applied to the solenoid coil E, provided the tuning assembly is not already in the position corresponding to the depressed pushbutton. The path of the current actuating the solenoid

is this: from the ungrounded terminal of the automobile battery, through terminals 8 of the connector (plug and receptacle), over the common lead (yellow) of the control head, through whichever button is depressed, through the connector, to one of the short contact fingers of the rotary selector switch. It may be seen in Fig. 5 that the contact ring of the switch has a cut-out, such that whichever finger is opposite the cut-out fails to make contact with the ring, whereas the circuit may be closed through any one of the other five contacts. If the button depressed corresponds to a finger of the switch making contact with the contact ring, current will flow from that finger through the ring, through the permanently connected finger, through the solenoid switch (which normally is closed) through the solenoid coil, and return through ground to the battery.

When current flows through the solenoid coil, E, the solenoid plunger is drawn into the solenoid coil-form by the magnetic forces developed. Since the plunger fits fairly tight in the coil form, the speed of this action can be controlled by the adjustable air release, which determines the rate at which the plunger can eject air from the coil form. When the plunger has been drawn in for nearly the full length of its travel, the rotating mechanism H turns the switch shaft (actuator rod) I one-sixth turn, thus turning the stop plate and switch by this amount. As may be seen at I, the switch shaft has a twist near the end within the solenoid; as the plunger draws the movable slotted cylinder H (ratchet) over this twisted portion one or the other must turn. The movable slotted cylinder is pressed against the fixed one by the action of a small coil spring inside the plunger, and the shape of the teeth on the two ratchets is such that the movable one can turn in only one direction. When the plunger is moving into the coil, the turning force on the ratchet is in the direction in which it will not turn; therefore, the switch shaft turns. However, when the plunger moves out, the turning force is in the other direction, and the ratchet turns, while the shaft is held in position by friction in the switch assembly.

Movement of the plunger out of the coil is produced by the action of the carriage return spring F. This spring is weaker than the magnetic pull of the solenoid coil, but the current through the coil is interrupted when the

plunger is fully inside the coil by the opening of the solenoid switch J. As the plunger moves out of the coil, the solenoid switch recloses. If the new position of the six-position switch B (it has just been turned 60° by the action of the slotted cylinders) is such that the short contact finger corresponding to the pushbutton depressed is not making contact with the contact ring of the switch, the solenoid coil circuit at E remains open despite the reclosing of the solenoid switch J. The spring is then free to pull the plunger out until the stop plate D is pressed against a lead screw stop G. However, if the circuit is closed through the rotating
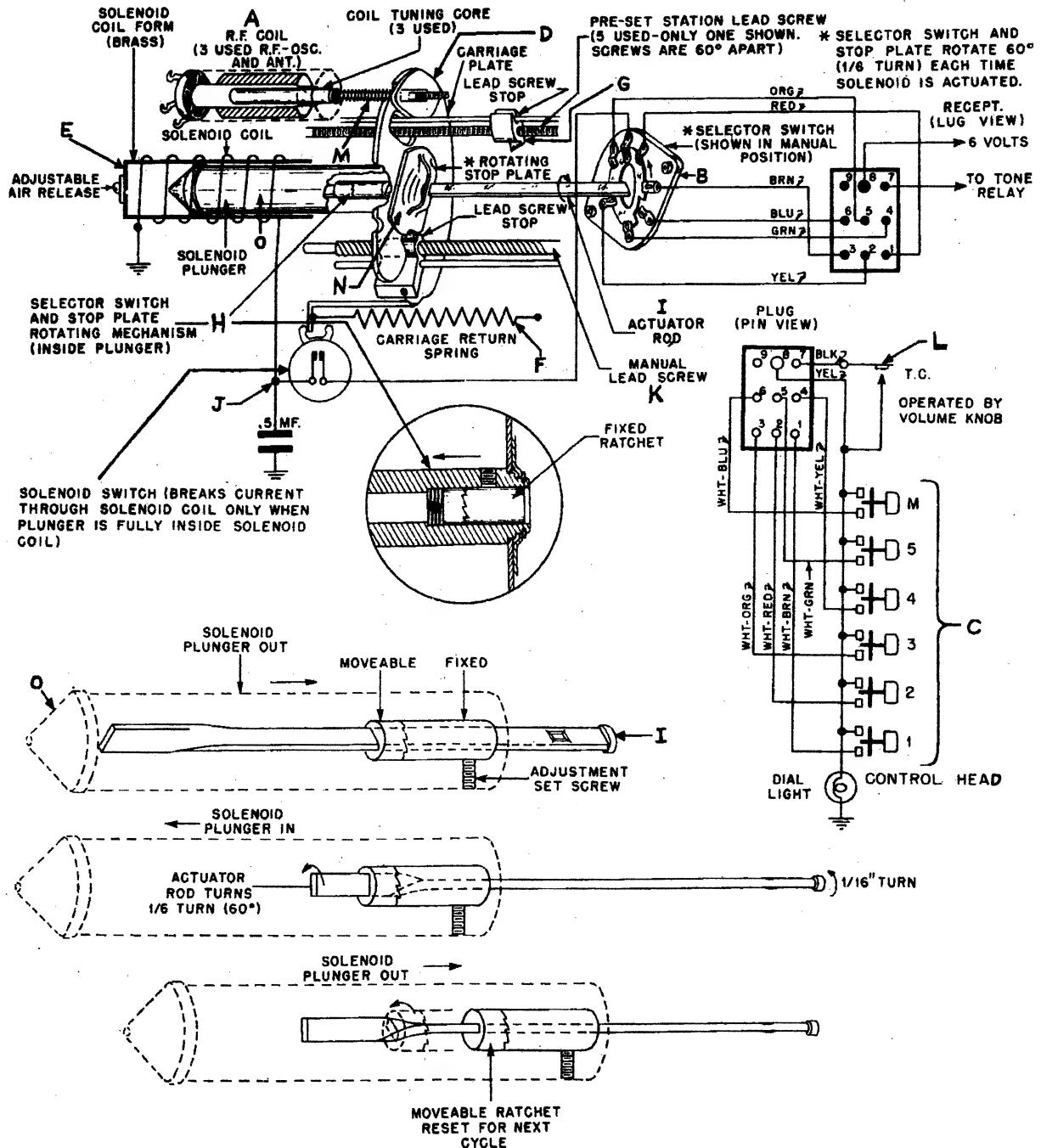
Fig. 5.—Inductive tuning is used in the Motorola solenoid-operated tuner Model ST-54 that is incorporated in several Galvin automobile receivers. Five pushbuttons provide means for selecting as many stations and a sixth permits manual tuning.

switch, the solenoid coil will pull the plunger back again as soon as the solenoid switch closes. The rotating switch is then turned another 60°; and the back-and-forth motion of the plunger, together with the turning of the rotating switch, continues until the rotating switch opens the coil circuit. Since there is only one position of the rotating switch, corresponding to each pushbutton, in which the circuit is opened at the rotating switch, pressing a button will cause the mechanism to seek and hold that one particular position which corresponds to the pushbutton pressed.

Tuning of the Receiver

Thus far we have not considered the actual tuning of the receiver. This is accomplished by varying the inductances of the r-f, oscillator, and antenna coils, while the tuning capacitors in the respective circuits are kept fixed. The variation of inductance is produced by moving the cores of the tuning coils, A and the motion of these cores is in turn produced by the carriage plate D to which they are attached by long screws M. The individual positions of the cores relative to the carriage plate may be adjusted as part of alignment, by turning the screws. The carriage plate is attached to the solenoid plunger, and thus moves with it.

Now to consider the positioning of the tuning cores for a particular station. Through the carriage plate pass six lead screws, see K, one for each pushbutton-tuned station and one for manual tuning. On each lead screw is a stop, G, whose position is determined by turning its lead screw, just as a nut moves along a bolt when the latter is turned. For each pushbutton-tuned station, the position of the corresponding stop is set when the pushbuttons are set up, while for manual tuning the sixth lead screw is turned by the manual tuning knob. The holes in the carriage plate through which the lead screws pass are large enough so that the stops may also pass through. However, mounted on the switch shaft or actuator rod, I, and turning with it, is a rotating stop plate, N, which

fits tightly against the carriage plate on the side away from the solenoid plunger, O. This stop plate N is so shaped that when no current flows through the solenoid, E, and the carriage return spring F pulls the carriage plate D toward the switch, B, it will strike against and be held in a fixed position by one, and only one of the lead screw stops, G. Which stop the stop plate rests against is determined by the position to which the rotary switch is turned.

As we saw above, the position of the switch is determined by which pushbutton is depressed; thus, the chain of events in tuning is this: When a pushbutton is pressed, the rotary switch B is caused to turn until it reaches a position corresponding to the pushbutton. In this position of the switch, the carriage plate D comes to rest in a position determined by the lead screw stop G corresponding to the pushbutton depressed. The position of the carriage plate in turn determines the positions of the three coil tuning cores, A, and thereby the frequency to which the receiver is tuned. Since the lead screw stops are adjustable, any five preset stations and as many manually selected stations as are within range of the radio may be tuned in.

On the control head there is a push switch L operated by depressing the volume-control knob. Closing this switch energizes the coil of the tone-control relay. When the relay is operated it pulls a hooked arm which engages a toothed wheel, giving it one-sixth of a turn. This wheel is on the shaft of a rotary switch, which has six fixed contacts and two movable contacts. The two movable contacts are opposite each other (that is, 180° apart on the switch rotor) and are connected together electrically. Therefore, in any position of the switch, two opposite fixed contacts are connected, so that there are only three different positions of the switch. Each position produces a different tone-control setting. Therefore, each time the volume-control knob is pushed, the tone control is advanced to a new setting, until three have been covered, after which the cycle is repeated.

CONTROL CIRCUITS

Control circuits in home and auto radios are mainly, except for the volume control, in the nature of extras. For example, many sets have no tone control, while some have a single tone control, which undoubtedly adds to the user's enjoyment of his set, and a few have both treble and bass controls, so that even greater flexibility in adjusting the set's characteristics to the demands of the individual listener is offered. Avc is a sort of border-line case; that is, it might be considered an extra, but it is so desirable that the great majority of radios include this feature.

Control circuits can be classified in two ways: as automatic and manual, or as a-f controls and r-f or i-f controls. In the former case, the controls are grouped in accordance with the agency which causes them to act; while in the latter the grouping is determined by what is controlled. In the discussion following, r-f and i-f controls are considered first. Several examples of special avc circuits and one manual i-f sensitivity control are explained. Under a-f controls both tone and volume controls are described, the latter including automatic volume expansion (ave) circuits as well as manual controls.

Chevrolet Model 985986

A screwdriver-adjusted sensitivity control is employed in the i-f amplifier of this receiver. This control is a variable resistance connected between the cathode of the 6SK7 i-f amplifier tube and ground, as shown in Fig. 1. Increas-
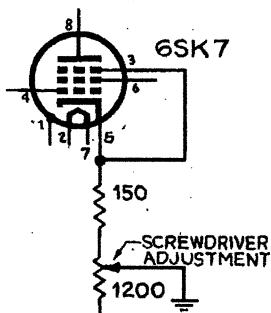
Fig. 1.—Grid-bias sensitivity control used in the Chevrolet Model 985986.

ing this resistance, increases the bias on the tube, thus lowering the gain; decreasing the resistance has the opposite effect.

The increase in bias with an increase in resistance is produced in this manner: For a certain value of cathode resistance, a certain cathode current will flow. The magnitude of this current will be just that necessary that its product with the cathode resistance (that is, the IR drop across the cathode resistance) equals the bias necessary for this current flow, with the particular values of plate and screen voltages applied to the tube. This can be stated as an equation: cathode current times cathode resistance equals bias.

If the resistance increases, the product of current and resistance (IR drop) will also increase; however, this increase is not as great as it would be if the current remained constant, for the current decreases, the decrease in current being caused by the increase in bias. Now it might be thought at first that the decrease in current might be sufficient to offset the increase in resistance, and thereby keep the IR product, which is the bias, constant. But if the bias were kept constant, so also would the current; therefore, the current must decrease, but by an amount insufficient to offset the increase in resistance. The bias, therefore, increases also when the resistance is increased, as was stated above. The bias decreases in the same manner for a decrease in the cathode resistance.

International Detrola Model 571X

A form of semi-fixed bias is used in this receiver on the avc-controlled grids. A diagram of the pertinent portions of the detector circuit is shown in Fig. 2. As a result of the so-called "contact potential", which exists in electron tubes, electrons flow from the cathode of the 14B6 to the diode plates even in the absence of a signal. The resultant voltage is applied to the avc-controlled grids of the preceding tubes.

When a very weak signal, insufficient to produce appreciable avc voltage, is received, the positive half cycles would be rectified if there were no bias on the grids. This would load the tuned circuits, reducing the sensitivity of the set. However, by applying a small bias this is avoided, while the reduction in gain due to the bias itself is very slight.

Unfortunately, some of the bias due to "contact potential" is applied also to the detector diode plate: that plate directly connected to the secondary of the second i-f transformer. This causes a slight lowering of sensitivity, but the effect is largely offset because only a small part of the bias applied to the avc-controlled grids is applied to the detector diode plate. The reason for this is two-fold: First, resistor R5
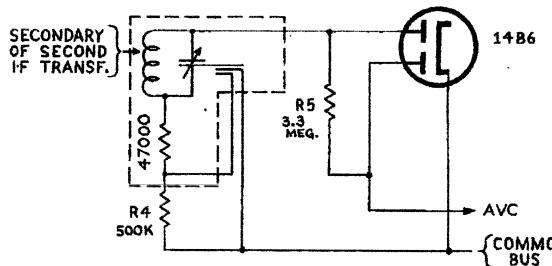
Fig. 2.—Semi-fixed bias on avc-controlled grids of preceding tubes in the International Detrola Model 571X.

(3.3 megohms), the 47,000-ohm filter resistor, and resistor R4 (500,000 ohms) form a voltage-divider (the resistance of the transformer secondary is negligible), with most of the voltage appearing across R5. In the second place, the "contact potential" effect due to the detector diode, is less than that due to the other diode, since the d-c load on the former is much less than that on the latter.

Pilot Model T-500

Although it is more or less general practice to use a common diode for both audio detection and avc rectification, in this receiver separate diodes (though with a common cathode) are employed. The part of the circuit under consideration is shown in Fig. 3.

The audio-detection diode plate is coupled
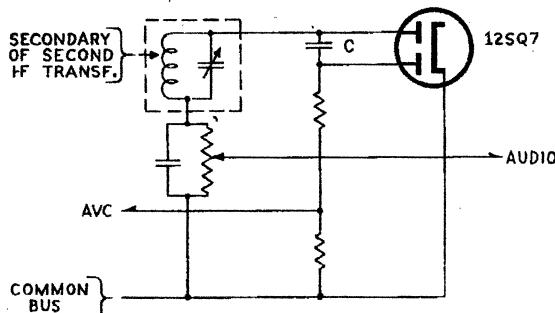
Fig. 3.—Avc and detector circuit used in the Pilot Model T-500.

directly to the secondary of the second i-f transformer, while the i-f signals are coupled from this winding to the avc rectifier diode plate through a capacitor. This capacitor C separates the d-c paths involved in the two circuits. This separation permits the choice of values for the resistors used that will be optimum for each application, that is, for audio and for avc. When a common diode is used, the choice of circuit values must be a compromise for best overall performance, and therefore, may produce poorer results than those obtainable with separate diodes.

Buick Model 980745

Space, of course, is at a premium in an auto receiver. In the Buick Model 980745 receiver it is conserved in the tone control-volume control system by mounting the controls one behind the other, as in a two-gang potentiometer. Separate shafts are provided and this is possible by using a hollow shaft for the tone control, which is mounted in front of the volume
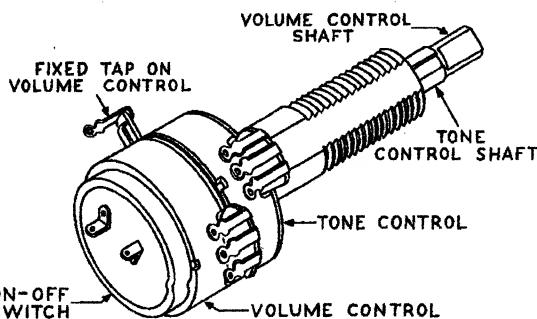
Fig. 4.—Combined manually operated volume and tone controls in the Buick Model 980745.

control. The shaft of the volume control passes through the hollow tone-control shaft. The on-off switch is ganged with the volume control, and is operated by the volume-control shaft in the usual manner. The mechanical features of this control, as described above, are illustrated in Fig. 4.

The volume control used here is of the tone-compensated type, which automatically changes the tonal characteristics of the radio as the setting of the volume control is changed manually. The reason for this compensation is that the frequency response of the human ear changes with the overall amplitude of the sounds heard. For example, if a certain piece of music is played loudly at first, then softly, it will *seem* to the listener that in the second

playing, the amplitudes of the high and low notes have been reduced more than the amplitude of the middle register, even though the amplitudes of all notes have actually been decreased by the same amount. For this reason, the volume control used here is compensated so that when the volume is turned down, the high notes are not decreased to the extent that the middle and low notes are.
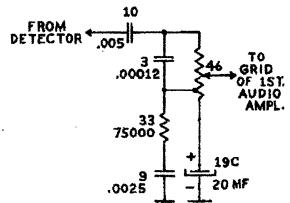
Fig. 5.—Schematic diagram of the volume-control section of the Buick Model 980745.

This effect is achieved by use of the circuit shown in Fig. 5. This diagram illustrates only the a-f portions of the circuit between the detector and the grid of the first audio amplifier; a few bypassed resistors and two bypass capacitors have been omitted for the sake of simplicity. Across the upper portion of the volume control (between the fixed tap Fig. 5 and the "loud" end) is a small capacitor, 3. This capacitor causes a relative lowering of the high-frequency audio (high notes) appearing across this portion of the volume control. The low end of the volume control can be considered to be grounded for audio frequencies, through capacitor 19C. Therefore, resistor 33 and capacitor 9 in series appear to be connected across the portion of the volume control between the fixed tap and the lower end. Although capacitor 9 has a greater value than capacitor 3, its shunting effect is much less because of series resistor 33. As a result, the proportion of high notes to middle and low notes appearing across the portion of the volume control between the fixed tap and the low end is greater than the proportion across the upper portion of the control. Thus, when the sliding tap is on the lower or softer, portion of the control, the high notes are *relatively* louder than when the tap is on the upper portion. This arrangement compensates, in part, for the non-linear characteristics of the human ear.

As may be seen from the above, the tonal properties of the volume control are not independent of the volume setting. For any given position of the volume control, a certain proportion of high, middle, and low notes will be passed, and this proportion is fixed for any one position of the control, although it changes as the position changes. Therefore, in order that

Fig. 6.—Schematic diagram of the tone-control section of the Buick Model 980745.
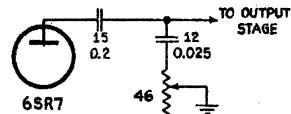
the tonal properties of the amplifier may be changed independently of the volume setting, a separate tone control is provided.

The tone control and closely related parts of the circuit are seen in Fig. 6. Capacitor 12 bypasses a large proportion of the high notes (high-frequency audio) to ground when the series resistance of variable resistor 46 is zero. When the full resistance of 46 is in the circuit, the bypassing effect of 12 is reduced very much, while in other positions of 46 the bypassing effect of 12 is set to intermediate values. In this manner, the proportion of high notes reaching the last stage can be varied over a considerable range by the setting of 46; thus, it serves as an effective tone control.

Chevrolet Model 985986 Tone Control

The tone control in this set is operated by a four-position, two-pole wafer switch, the complete circuit of the control and "*clarified schematics*" being shown in Fig. 7A. In the "Bass" position Fig. 7B, capacitor 34 bypasses a large proportion of the treble notes (high-frequency a.f.) to ground, thus accentuating the bass notes by comparison. In the "Soft" position Fig. 7C, capacitor 33 replaces capacitor 34; the circuit is otherwise unchanged, but since the capacitance of 33 is less than that of 34, the reduction of the treble notes is not so pronounced as in the "Bass" position of the tone switch. When the tone switch is thrown to the "Music" position, Fig. 7D, a direct connection is made from the first audio stage to the second.

In the "Voice" position both high and low notes are attenuated. This, of course, makes human speech sound somewhat less natural than when the treble and bass are reproduced in the same relationship as in the original sound. However, it has been found experimentally that in the presence of extraneous noises, such as wind rushing past an automobile, traffic, or static, etc., etc. the intelligibility of human speech is increased by accentuating the middle frequencies. This occurs because most of the audio

power in speech is in the middle frequencies and because the ear is most sensitive to the middle frequencies.

In this receiver the accentuation of the middle frequencies is produced by reducing the

7 (A)

7(B)
BASS

7(C)
SOFT

7(D)
MUSIC

7(E)
VOICE

7(F)
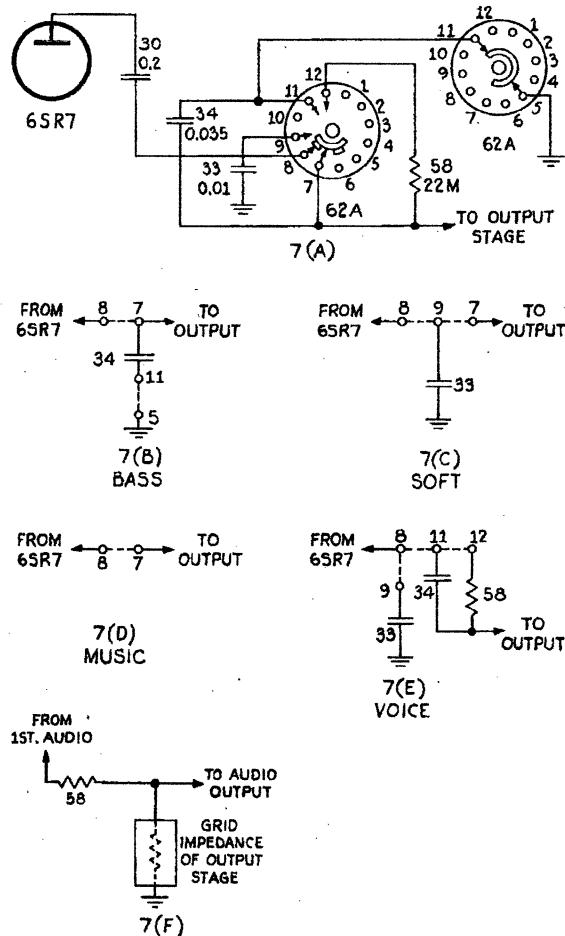
Fig. 7.—The four-position two-pole tone-control switch used in the Chevrolet Model 985986, is shown in Fig. 7(A). The various components affected by the switching are shown in the other special schematics.

high- and low-frequency portions of the audio signal by the components shown in Fig. 7E. The treble is cut by capacitor *33*, which bypasses part of it to ground. The bass is lowered by the parallel combination of resistor *58* and capacitor *34*. Momentarily disregarding the effect of capacitor *34*, resistor *58* and the grid impedance of the output stage form a voltage divider across the first audio, which is shown in Fig. 7F. Because of this voltage divider only part of the output of the first stage is transferred to the output stage. This is completely true at low frequencies, in which region the shunting effect of capacitor *34* on resistor *58* is negligible; how-

ever, at middle and high frequencies Fig. 7F is incomplete, for the shunting effect of capacitor *34* shown in Fig. 7E, becomes important. At these frequencies, then, the voltage-dividing effect is negligible, because the parallel impedance of capacitor *34* and resistor *58* is very small. Thus, attenuation of both high and low frequencies is obtained, the former by means of capacitor *33*, the latter by means of the combination of capacitor *34* and resistor *58*.

Crosley Model 66CA

The volume and tone controls in this set are interconnected very closely, as shown in Fig. 8A. At first glance it might appear that the tone control is a feedback circuit; however, if Fig. 8A is redrawn, as in Fig. 8B, it is seen that although potentiometer *44* functions in both the grid and plate circuits of the 6SQ7, it does not provide feedback, for it can be separated into two independent parts, connected only through ground.

The volume control *45* is of the tone-compensated type. In this case, however, compensation is in favor of the bass notes, since the effect of resistor *15* and capacitor *34* (omit tone control *44* temporarily), which are connected from the tap on the volume control to ground, is to reduce the treble notes at low settings of the volume control. But when the tone control is put into use, the reduction of high notes is variable, for the bypassing effect of capacitor *34* may be reduced to zero if the movable arm of *44* is turned to the end to which *34* is connected. Doing this shorts out *34*, for both ends are now grounded.

What is the effect of the tone control in the plate circuit of the 6SQ7? Capacitor *25* tends to bypass the high audio frequency, but the extent to which this actually occurs is controlled by the amount of resistance inserted by control *44* between *25* and ground. When the arm of *44* is turned to the position where capacitor *34* is shorted, the resistance between capacitor *25* and ground is maximum, and the greatest available high-frequency response is obtained in both the grid and plate circuits. On the other hand, when the arm of the tone control is turned to the end to which capacitor *25* is connected, the bypassing function of *25* is effective to the fullest extent. At the same time, capacitor *34*, instead of being shorted, is shunted by a high re-
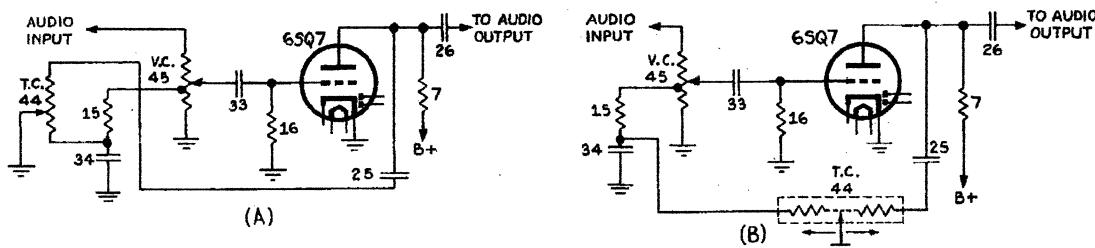
FIG. 8.—In Fig. 8(A) are shown the tone-control and volume-control circuits of the Crosley Model 66CA; the equivalent circuits are shown in Fig. 8(B).

sistance, which has little effect. Thus the treble notes are considerably reduced in both the grid and plate circuits.

Motorola Model CR6

A volume and tone control system similar to that used in the Crosley Model 66CA is found in this set. Fig. 9 shows the circuit of these controls. (This diagram is complete with regard to the tone and volume controls *only*).
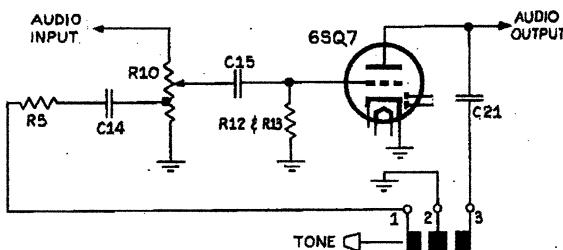
FIG. 9.—The three-position tone-control circuit and volume-control circuit in the Motorola Model CR6.

The tone switch has three positions. In the "Voice" position all contacts are open; *R10* functions as an uncompensated volume control and *C21* has no effect, since the switch ends of *R5* and *C21* are open. The second position, "Music," provides bass compensation of the volume control by connecting *R5* to ground, contacts *2* and *3* of the switch remaining open. In the "Bass" position, all contacts of the switch are closed; thus the volume control is compensated in favor of the bass notes, and the treble notes are attenuated in the plate circuit by *C21*.

Goodrich Model 664PM

A doubly tone-compensated volume control is used in this receiver to produce a more natural bass accentuation by attenuating the highs at low-volume levels than can be obtained with a single tap. Fig. 10 shows the connections to the volume control to obtain double compensa-

tion. In the usual compensated control, a single fixed tap is used. This produces a rather abrupt change when the movable tap on the volume control is rotated past it. The human ear is not so sensitive as to notice this as an objectionable

FIG. 10. — Doubly compensated volume control in the Goodrich Model 664PM.
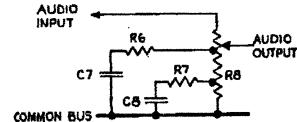
effect; the contrary is true, or this type of compensation would not be used. However, it is not altogether natural. By using two steps of compensation, each of which alone produces a smaller change in the tonal characteristics of the receiver than a single one, a more gradual variation in bass accentuation is obtained. Thus the effect of naturalness obtained by compensation is hightened by the use of doubly compensated control.

Hoffman Model A 501

Both bass and treble tone controls, and a tone-control tube, which isolates the controls from each other, are used in this receiver. The circuit of this tone-control stage is shown in Fig. 11. Both controls function by varying the gain of the stage.
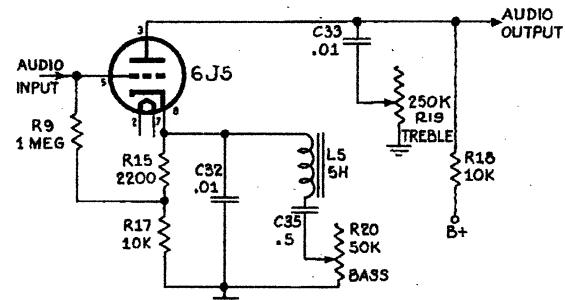
FIG. 11.—Schematic diagram of the bass and treble tone controls and the tone-control tube in the Hoffman Model A501.

A large cathode resistance in an amplifying stage causes considerable degeneration; that is, a reduction in gain relative to that available when the cathode resistance (or, more exactly, impedance) is reduced effectively to zero. The cathode impedance in this stage is a rather complicated combination of capacitors, resistors, and a choke. At very low frequencies (below the audible range) C32 and C35 have very high reactances, so that the effective cathode impedance is equal to R15 and R17. This results in quite a low stage gain. At somewhat higher frequencies, L5 and C35 approach series resonance, which they reach at 100 cycles. At this frequency, their series impedance is practically zero, and if the bass control R20, is reduced to zero also, R15 and R17 will be bypassed completely. In this case, the stage gain will be relatively high at 100 cycles, and even at 50 and 200 cycles (one-half and double the resonant frequency) this series-tuned circuit produces a considerable rise in gain if the resistance of R20 is cut to a low value. If the resistance of R20 is made high, however, the effectiveness of the series-tuned circuit in bypassing R15 and R17 is reduced considerably, and the rise in gain for bass notes is likewise reduced. Thus R20 enables the used of the receiver to vary the bass boost from a very small amount up to a considerable amount (the range is approximately 8 decibels). C32 provides partial bypassing at high audio frequencies (at 8000 cycles, its reactance is approximately 2000 ohms). This produces a slight rise in gain at high frequencies.

The treble control provides a variable high-frequency bypass across the load resistor, R18; in this manner the high-frequency response can

Fig. 12.—Treble and bass tone controls are used in the Magnavox Chassis CR-187. The bass control employs positive feedback so that the effective gain of the inverter is increased for bass notes.

be varied. If R19 is at its maximum value, the bypassing effect of C33 is negligible, and treble boost will be obtained by the action of C32 in the cathode circuit, as explained above. At the other extreme (R19 reduced to zero), the bypassing effect of C33 will be so pronounced as to more than overcome the boosting effect of C32, and a considerable degree of treble cut will be obtained. The range of control at 8000 cycles is about 16 decibels.

The purpose of the tap on the cathode resistors of the tube (at the junction of R15 and R17) is to provide a suitable bias voltage for the grid.

Magnavox Chassis CR-187

Two tone controls are employed in this receiver: one for treble and one for bass. The treble control is quite simple. Fig. 12, which is a simplified schematic of the audio inverter and output stages, shows that the treble control in effect, is a variable bypass for high-frequency audio. Bypassing is provided by capacitor 23, while its effect is varied by treble control 67. This control affects the output tubes equally, for the bypassing effect applies to the signal fed through the lower half of the inverter (6SN7) and so to both the lower and upper 6V6 tubes.

The bass control, on the other hand, operates in a relatively complicated manner. Briefly, it uses positive feedback over a path which passes the low frequencies predominantly, so that the effective gain of the inverter stage is increased for bass notes.

Positive feedback is obtained in this manner: The 6SN7 operates as a conventional inverter, with the drive to the lower grid (pin 4) derived from a tap on a voltage divider connected across the output from the upper plate (pin 2). The output from the lower plate (pin 5) is therefore in phase with the input to the upper grid (pin 1). A portion of this output is fed back through resistor 56 and capacitor 15 to the low end of the return resistor 49 of the upper grid (pin 1).

Since this voltage is in phase with the input signal from the preceding stage, it adds to the signal, increasing its effective amplitude. This increase in the effective amplitude of the input produces a proportional increase in the output. Although the coupling capacitor 15 in the feedback path is small, the signal fed back
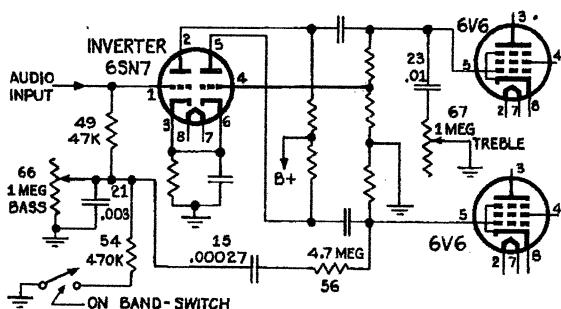
consists predominantly of low-frequency voltages, since the middle- and high-frequency components are bypassed to ground by capacitor *21*. Therefore, the increase in output due to feedback is confined principally to the lower frequencies; in this manner bass accentuation is obtained. Because control *66* is variable, the proportion of the output voltage that is fed back can be varied, and in this way the bass accentuation is controlled. When the band switch is in the short-wave position, resistor *54* is placed in parallel with the lower arm (capacitor *21* and that part of control *66* in actual use) of the voltage divider in the feedback path. This reduces the maximum percentage of feedback available, and therefore, the maximum bass accentuation that can be obtained on the short-wave band.

Packard-Bell Model 1052

Two interesting and unusual features in this receiver are a tone-compensated volume con-
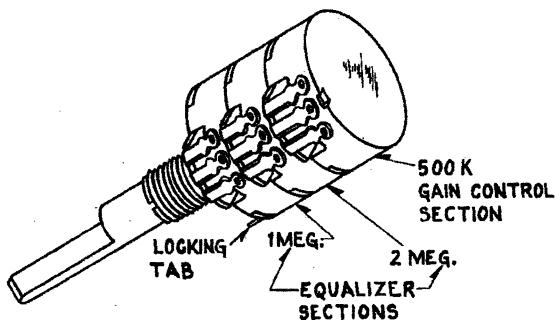

FIG. 13.—The special three-gang potentiometer used as a tone-compensated volume control in the Packard Bell Model 1052.

trol, using a three-gang potentiometer instead of the usual single potentiometer with a fixed tap, and an automatic volume expander. The special three-gang potentiometer is illustrated in Fig. 13, and a schematic diagram of the essential parts of the circuit associated with it, is shown in Fig. 14.

The purpose of tone compensation in this volume-control circuit is the same as in any other, that is, to compensate for the varying characteristic of the human ear. As was pointed out previously, the use of a volume control with a fixed tap and a fixed equalizing network connected across part of the volume control, produces a desirable compensating effect. (An equalizing network is a combination of resis-

tors, capacitors, and/or chokes which has a certain desired frequency response. In the case of the tone-compensated volume controls discussed in this section, resistor-capacitor combinations are used to give an increased response to bass notes when the volume is turned down.) However, this effect is rather abrupt, that is, when the movable arm on the volume control rotates past the fixed tap, a sudden change in the tonal characteristics of the audio amplifier takes place. Although it is a considerable improvement over no compensation at all, it does not produce an altogether natural effect.

By using variable equalization, rather than fixed, a tone-compensation characteristic can be obtained that changes gradually with the volume-control setting and comes very close to compensating exactly for the peculiarities of what might be termed the "average" human ear. Variable equalization is produced in the Packard-Bell Model 1052 by using two variable resistors in the equalizing network, and ganging them with the volume control. The volume control is *R19*, while the ganged variable resistors in the equalizing network are *R20* and *R22*. In addition to the degree of equalization determined by *R20* and *R22*, and dependent on the volume-control setting, the listener has at his disposal a separate tone control, *R23*. Being part of the equalizing network, *R23* can vary its characteristics to suit the individual listener.

This equalizing circuit operates from 50 to 400 cycles. With the volume control turned down to the point where only 100 milliwatts of audio power are supplied to the speaker, the bass accentuation at 50 cycles is 25 decibels with *R23* in the maximum bass position. With *R23* in the maximum treble position, the bass
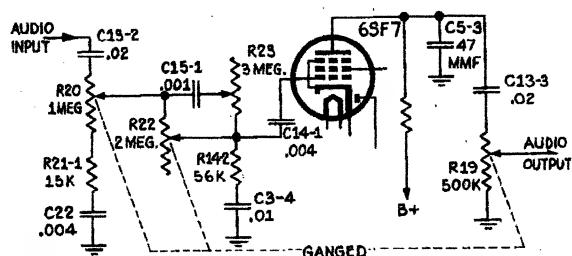

FIG. 14.—How the ganged volume control shown in Fig. 13 is connected in the first a-f amplifier.

boost is reduced to 10 decibels for the same volume-control setting. The equalizing circuit

is so designed that there is no bass boost when the volume control is turned to the high end past the half-way point. At this point, the tone control also ceases to become effective.

Automatic Volume Expansion

In the article on "Recording" the use of automatic volume compression in the Packard-Bell Model 1052 is discussed. As was pointed out there, no volume control is necessary in the recording circuit because the gain of the amplifier is held automatically at the proper value. This results in well-cut grooves on the records, but the volume range of the recorded sounds is so confined by compression as to sound somewhat unnatural. This condition is remedied by the automatic volume expander (ave), which has just the opposite effect of the compression circuit; that is, variations in volume of the sound are accentuated by the ave, instead of being suppressed, as in the compression circuit.

The automatic volume expander (ave) includes two features not ordinarily found in conjunction with this type of circuit. These are delayed audio avc and a limiting diode. The former opposes the ave action at high levels, so as to cause a tapering-off effect and so prevent overloading. At low levels, the delay prevents audio avc operation, and the ave operates unaffected. The limiting diode has much the same function as the delayed avc. The avc operates when high levels are maintained over a relatively long period, but the time constants of its

Fig. 15.—The automatic volume-expansion circuit of the Packard Bell Model 1052.

circuits are too long for it to operate on sharp peaks; in such a case, overloading is prevented by the limiting diode.

The ave circuit is shown in Fig. 15. Audio voltage from the phono pick-up or second detector is fed to the volume-control equalizer and also to the control grid of the expansion amplifier. Resistors R16-2, R16-4, and capacitor C8-6 perform a decoupling function. They prevent negative audio feedback from the plate of the expansion amplifier to its control grid. At the same time, these resistors provide a d-c path from one of the plates (pin 3) of V9 to the control grid of V6. In addition, these resistors, together with R16-1, form the ground-return path for the control grid, and a bias voltage is developed over them by the "contact potential" effect between cathode and grid of V6.

At low audio levels neither the delayed audio avc nor the limiting diode operate. The ave action, then, is this: The a-f signal, amplified by V6, is fed through C8-4 to a plate (pin 5) of V9. A positive d-c (rectified) voltage proportional to this audio appears across R16-5, and is applied through R15-3 and R16-7 to the control grid of the first audio amplifier, V7. Since the cathode of V7 is connected to a voltage divider across the B+ supply, this puts it at a potential of 5.9 volts above ground. The rectified voltage from V9 reduces the grid-cathode bias on V7. This in turn increases the gain of the first audio stage. Thus the gain of V7 is made proportional to the average level of the audio signal, and changes in the average level are accentuated, for relatively high gain is provided for high signals and low gain for low signals. The a-f signal is fed to the control grid of V7 through C14-1 from the volume-control equalizer. In conjunction with this, resistor R16-7 and capacitor C17 decouple the audio from V9; in addition, C17 acts with R15-3 as a filter in the output of V9.

At high audio levels that are reached abruptly (that is, sharp peaks), the rectified voltage applied by V9 to the control grid of V7 may tend to exceed 5.9 volts. Since this is the cathode voltage of V7, the grid would go positive in the absence of a limiting device. Should the grid go positive, considerable audio distortion would be produced. The limiting device is the diode plate of V7. Any voltage from V9 in excess of 5.9 volts will cause current to flow from the cathode of V7 to the diode plate of that tube. This cur-

rent will produce a voltage drop across $R15$-3 which will buck out the excess voltage, thus preventing the control grid from being driven positive.
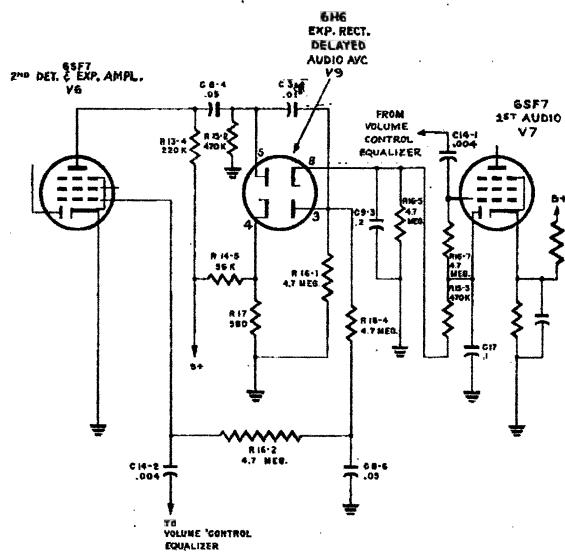
When a high audio level is reached gradually, the delayed audio avc operates. The voltage delay is produced by a fixed bias applied to the cathode (pin 4 of V9) of the avc diode by the voltage divider $R14$-5 and $R17$ from B+ to ground. This fixed bias is 1.55 volts. The amplified audio signal from the expansion amplifier, V6, is fed through $C8$-4 and $C3$-4 to the plate (pin 3) of the avc diode. If this a-f signal is higher than the delay bias, it is rectified, and the resultant negative d-c bias is applied to the control grid of V6. This negative bias decreases the gain of V6, and therefore reduces somewhat the output of the expansion rectifier (pins 5 and 8 of V9). In this way, the amount of expansion is reduced at high levels, so as to prevent overloading. The time constant of the path from pin 3 of V9 to the control grid of V6 is such that the action of the delayed avc does not commence immediately after the input to the delay diode exceeds the delay bias. For this reason, the limiting diode is occasionally required to act.

Philco Model 46-1201

In this radio-phonograph combination, a single knob operates the volume control, on-off switch, and radio-phono switch. The on-off switch is ganged to the volume control in a manner similar to that usually found when these two controls are ganged. The difference lies in the unusual "OFF" position, which occurs when the volume control is at the center of its rotation. The reason for this is that the receiver operates when the knob is turned to the right (clockwise) from the central position, and the phonograph when the knob is turned to the left (counterclockwise).

A further unusual feature necessitated by this arrangement is the grounded center tap on the volume control. This is required because zero volume is found at the central (off) position. Receiver volume is on full when the knob is turned all the way to the right, and phonograph volume is loudest with the knob turned as far as it will go to the left. The radio-phono switch (which controls power to the turntable, etc.) is operated by a cam on the shaft of the volume control. Thus three functions are controlled by a single knob.

OSCILLATOR-BIAS CONTROL

One of the important characteristics required of a frequency-conversion system is high mixer-stage gain. This characteristic of a mixer or converter is sometimes called translation gain or more usually *conversion gain*. It is equal to the ratio of the intermediate-frequency voltage at the output of the mixer to the radio-frequency voltage at the input of the mixer.

Mathematically this is seen as:—

$$Conversion\ gain = \frac{mixer\ output\ i\text{-}f\ voltage}{mixer\ input\ r\text{-}f\ voltage}$$

It could be shown that conversion gain is also equal to the product of the total load impedance of the tube multiplied by the *conversion transconductance* of the tube. This latter term may be defined as the ratio between the intermediate-frequency current at the output of the mixer to the radio-frequency voltage at the input of the mixer, and is given the symbol G_c.

Mathematically this is:

$$G_c = \frac{mixer\ output\ i\text{-}f\ current}{mixer\ input\ r\text{-}f\ voltage}$$

Conversion transconductance is a very important quality of a mixer tube since the gain of the stage depends upon having a high value of G_c.

The conversion transconductance of a mixer (and converter) tube is an indirect function of the oscillator voltage injected into the mixer tube, whether this is done through the cathode, control grid (first grid), or suppressor (third grid). The oscillator voltage injected into the mixer is quite large compared to the input r-f signal voltage so that the oscillator voltage controls the amount of plate-current flow in the mixer.

The voltage output of a conventional high-frequency heterodyne oscillator in use in a superheterodyne receiver, varies somewhat with frequency. Thus the voltage output of the oscillator is different at one end of the tuning range than it is at the center or at the other end of the tuning range. Whether the oscillator output increases or decreases as the oscillator frequency goes up, depends on the impedance relations of the oscillator circuit. This latter point does not require elaboration for the purposes of this discussion.

Since the output of the oscillator determines indirectly the conversion transconductance of the mixer tube and thus the conversion gain, changes in the oscillator output cause the conversion gain of the mixer to vary with frequency. This is, of course, a disadvantage in receivers, since overall receiver gain will tend to vary at different points in the tuning range. In addition, sensitivity will not be uniform throughout the frequency range of the receiver.

Maximum gain of the mixer stage at the conversion frequency (intermediate frequency) is obtained when the mixer tube is operated at the peak of its "conversion-transconductance versus grid-bias curve," which usually is called "conversion-transconductance curve." Fig. 16 illustrates a typical conversion-transconductance versus grid-bias curve and shows how the conversion transconductance of the mixer varies with the bias on the tube. To obtain the optimum operating condition, the tube must be operated at a definite bias. The normal procedure when designing the receiver mixer circuit is to set the mixer bias at a median point in the receiver frequency range so as to operate the mixer at the peak of its transconductance curve.

When a mixer tube is operated at the peak of its conversion-transconductance curve for a particular oscillator frequency (usually a median frequency), variation in the oscillator frequency (with receiver tuning) will vary the oscillator voltage output, as explained above,
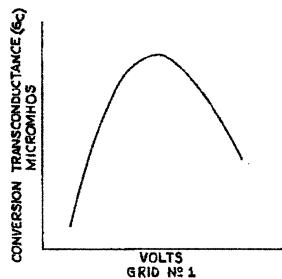

Fig. 16.—Curve showing how the conversion transconductance varies with the grid bias of the mixer tube.

and will shift the point of operation of the mixer to one or another side of its G_c peak.

This effect can be anticipated by making use of an oscillator whose voltage output is constant regardless of frequency. This, however, is not an easy or inexpensive solution and it is necessary to look elsewhere for solving this problem.

Since variation in tube bias will change the tube G_c, anything which will vary the tube bias will at the same time vary the tube G_c. The

discussion below will indicate how use is made of this principle to solve the problem of variation in conversion gain.

Sears Roebuck Model 6092

A circuit arrangement enabling constant operation of the mixer tube at the peak of its conversion-transconductance curve and thus guaranteeing maximum mixer-stage gain (conversion gain) is used in the Sears Roebuck Models 6092, 6093, 6104A, etc. The portion of the schematic of the first two models, relative to this discussion, is illustrated in Fig. 17.

These receivers make use of an oscillator bias-control circuit to accomplish the above condition. The oscillator bias-control tube is a 7E6, which consists of two diodes and a medium-mu triode in a single envelope. One diode section of the 7E6 is not used, its anode being grounded, while the other is used to develop a supplementary bias voltage for the mixer tube (here termed a translator). The bias voltage is fed into the signal-grid input circuit of the mixer. The oscillator section of the 7E6 feeds the oscillator voltage from its cathode into the cathode of the mixer tube. The diodes in the 7E6 are well shielded from the triode section so that there is freedom from interelectrode interaction between the respective circuits.

The mixer (translator) is a 7H7 tube whose characteristics are similar to the 6SG7 tube. The 7H7 is a pentode super-control voltage amplifier having high transconductance. The tube may further be described as a *semi*-remote cutoff tube and the reason for its use here may be described briefly as follows:

One of the features desired in a mixer is control of amplification by variable grid bias. This requires that a gradual cutoff tube be used. However, high conversion gain is also desired and this requires the use of a sharp cutoff tube. The semi-remote cutoff tube is thus a compromise between these two requirements.

The circuit shown in Fig. 17 creates a bias voltage for the mixer tube dependent on the oscillator output. This may be seen from the fact that the 7E6 cathode is common to both the triode oscillator circuit and the bias-control diode circuit. Changes in mixer bias due to changes in oscillator output take place as follows:

When the oscillator output increases, the voltage drop across the oscillator output load
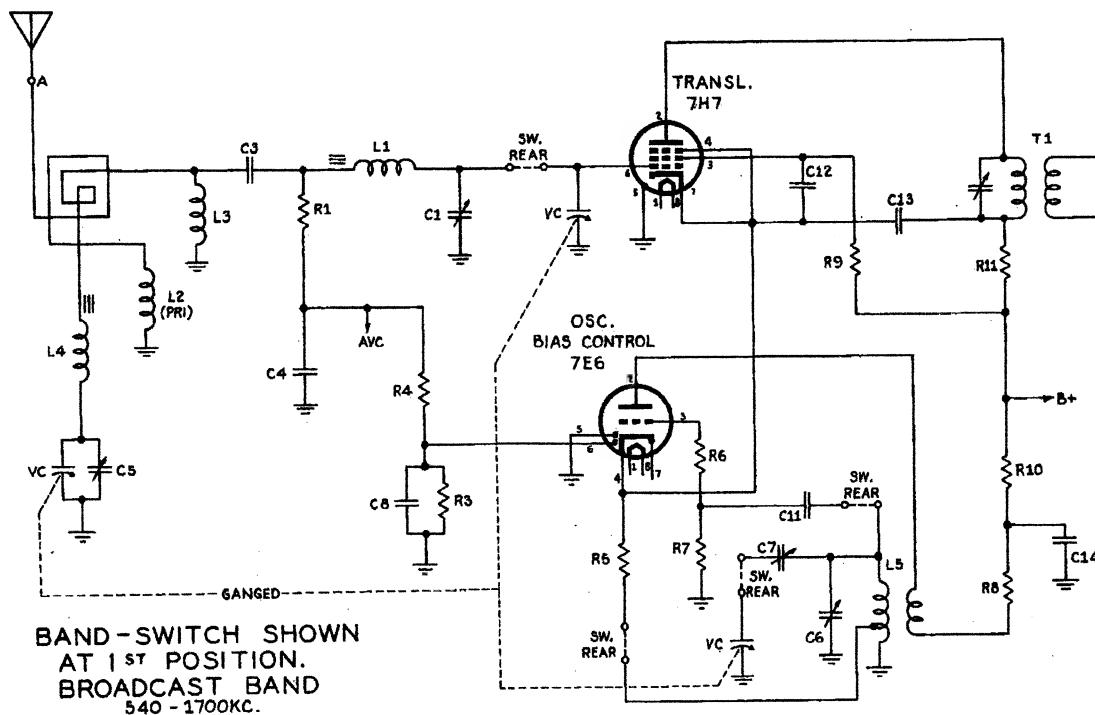
Fig. 17.—Schematic diagram showing how constant operation of the mixer tube at the peak of its conversion-transconductance curve is obtained in the Sears Roebuck Model 6092.

increases, thus increasing the oscillator plate current. The oscillator plate current passes through the cathode circuit, and this current, whose drop through the cathode resistance is used to provide oscillator voltages for the mixer tube, is also rectified by one of the 7E6 diodes. Resistor $R3$ (1 megohm) and capacitor $C8$ (100 mmf) serve as an r-f filter to eliminate the oscillatory components in the rectified voltage. Resistor $R4$ (1 megohm) serves to decouple the bias-control circuit from the normal avc circuit.

The voltage developed by the bias-control diode is impressed across the mixer grid load and serves to augment the self bias developed by the mixer cathode resistance. Thus the bias on the mixer will vary with changes in amplitude of the oscillator output and this will com-

pensate for the shift in mixer conversion transconductance resulting from changes in oscillator voltage. The regulation of the mixer bias by the oscillator cathode current causes the mixer to swing back automatically to peak operation each time a change in oscillator voltage causes it to swing away from peak operation. If the oscillator output increases, the bias on the mixer tube will increase and if the oscillator output decreases, the bias will decrease, so that the tube is operated at the peak of its conversion-transconductance curve at all times.

Maintenance of the conversion gain constant in this manner permits normal operation of the avc which is fed to the mixer and i-f amplifier from the 6SQ7GT tube.

"GIMMICKS"

The title heading of this page will probably arouse the curiosity of many a reader. The word, gimmick, has been used so widely throughout the radio industry (besides being used in ordinary everyday life) that an exact meaning of the word becomes difficult. We have found reference given to gimmicks by many receiver manufacturers and analyzing all the cases, we have set up an arbitrary definition of what a gimmick is. *Any piece of wire or coil that has been inserted purposely in a circuit in such a manner that one end is connected in the circuit with the other end open, or both ends open, and serves the purpose of a small capacitance, is termed a "gimmick".* In other words, a certain amount of capacitance exists between the open end of the wire or coil and some other part of the circuit.

An illustration will help explain the full extent of a gimmick. In Fig. 1A is shown a coil of wire $L1$ with one end tied to the control grid of a tube and the other end open. Near this open end of the coil there is an inductance $L2$. The inductance $L2$ is part of a 1000-kc resonant tank circuit of which $C2$ is the variable capacitor and the tank circuit is excited by a 1000-kc signal. Due to this excitation, a circulating alternating current exists within the tank circuit and energy is available from this

tank. Placed very close to the inductance of the tank is the open end of $L1$, which forms, effectively, one "plate" of a capacitor. The high side of $L2$ represents the other "plate" of a capacitor and together both "plates" form an effective capacitance, which is small in value, some of the energy from the tank circuit being fed into the grid of the tube through the $L1$ gimmick.

An equivalent circuit with the gimmick replaced by a capacitor is shown in Fig. 1B. Since it may not be evident exactly how a capacity is formed between $L1$ and $L2$, it will be best to study the formation of a capacity from the physical point of view.

Two conductors separated by some type of dielectric can be called a capacitor; the dielectric may be air, paper, mica, oil, or a number of other materials. The many different insulating materials used as a dielectric between two conducting plates are capable of exhibiting electric charges on the outer surface of the material facing the conducting plates. How capable these insulating materials are in exhibiting electric charges is dependent upon the type of dielectric used.

A *standard* has been set up to enable us to have a method of determining the extent of how well these materials will exhibit electric charges. The term *dielectric constant* or *specific inductive capacity* is used as a reference of the ability of the dielectric to exhibit electric charges and the dielectric of a vacuum is taken as the standard and has a dielectric constant of unity (1). For most general purposes and for quick calculations, air has been considered as the standard most often used and it also has been given a dielectric constant of about one (1). (The actual difference between the dielectric constant of air and a vacuum is so small that for most purposes it is considered as negligible). All the dielectric constants of the other insulating material used are based upon the dielectric constant of air which is unity. When we refer to the *dielectric constant* of any material used as a dielectric insulator (other than air) between any two conducting plates we mean the following:
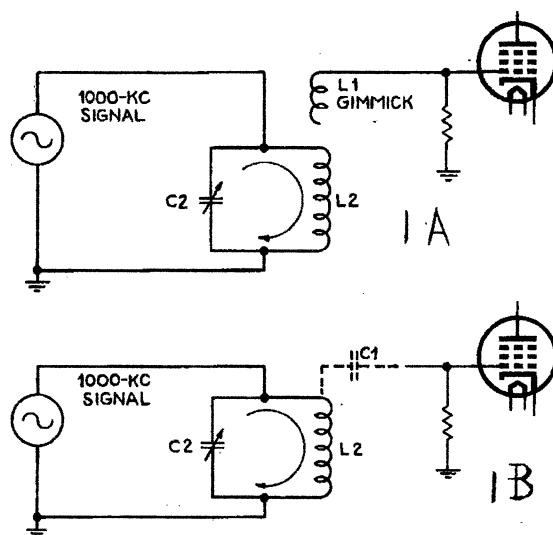

Fig. 1.—Illustrating how a gimmick functions as a capacity in a circuit.

The *ratio* of the true capacitance between the plates when using any dielectric other than air, to the capacitance between the same plates when air is used as the dielectric, is termed the *dielectric constant* of the insulating material.

The capacitance between two conducting plates can be calculated if the distance between the conductors, the common area between their plates, and the dielectric constant of the insulating material used are known. It is not intended here to determine the actual capacitance between such a system as in Fig. 1, but rather to show that a capacitance does exist between the coils. To be more exact in the calculation of a capacitance between two conductors separated by a dielectric, we will consider the simple case of two flat plates separated by air. For all general purposes, the capacitance of any two parallel-plate conductors is given by the following formula:

$$C = .225 \frac{KA}{d} \text{ in mmf}$$

where

K equals the dielectric constant of the insulating material

A equals the common area between the plates in square inches

d equals the distance between the plates in inches

Therefore, if the area common to the plates is 3 square inches, the distance between the plates 0.1 inch, and the dielectric constant equal to 1 for air then the capacity would be:

$$C = .225 \frac{1 \times 3}{0.1} = .225 \times 30 = 6.75 \text{ mmf}$$
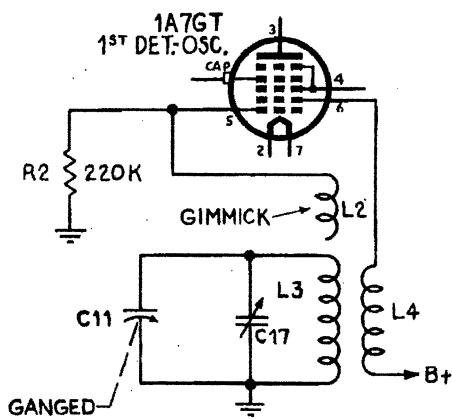
Although the coils under discussion do not in any way represent a parallel-plate capacitor, the above problem will illustrate the nature of how a simple parallel-plate capacitor is determined and that capacitance does exist between $L1$ and $L2$ of Fig. 1 as described. The actual calculation of the capacity between $L1$ and $L2$ is in itself too complex to compute, but it is much easier to measure it with some suitable measuring instrument such as a Q meter.

With the understanding of how a capacity does exist and of what is arbitrarily defined as a "gimmick", the next best thing to do is to study the various different receivers that use one form of gimmick or another and to understand the manufacturer's reason for using his gimmick.

RCA Model 55F and Firestone Model 7398-9

In both the RCA Model 55F and Firestone Model 7398-9 a gimmick is used in their oscillator circuits to provide the necessary capacitance to couple the oscillator-tank-circuit voltage to the grid of the oscillator. The oscillator circuits of both these models are illustrated in Fig. 2. Both models have a form of tickler-coil oscillator but seem to have their tank circuit "floating". That is to say, the oscillatory tank circuit has only one end going to ground and the other end unconnected. At a first glance one would believe that there is no means of conductively coupling the oscillator-tank voltage to the grid of the oscillator section of the tube. Now that something is known
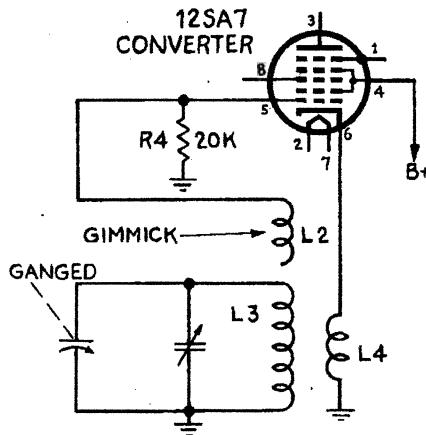

Fig. 2.—At the left is shown the RCA Model 55F, and at the right is shown the Firestone Model 7398-9; both using a gimmick *L2* in the oscillator circuit.

about gimmicks, notice the coil *L2* placed above the tank inductance *L3;* hence an effective capacitance exists between the open end of *L2* gimmick and coil *L3* so that some oscillatory voltage is fed through this effective capacitance, formed by the gimmick and high side of *L3*, to the grid of the oscillator tube. These gimmicks usually are wound around the same coil form as the rest of the oscillator coils and how close they are wound or coupled to the other coils determines the amount of capacity introduced between the gimmick and tank coil and hence the *amount* of oscillatory voltage fed to the grid.

This kind of a gimmick replaces the usual small mica coupling capacitor that is often used to couple the oscillator voltage to the grid of the oscillator tube. With such an arrangement, much needed space is saved and that is always required in the manufacture of radio receivers. The serviceman will probably not be able to notice from looking at the chassis that a gimmick does exist in the oscillator circuit since most of these gimmicks are on one oscillator coil form and the form is completely wax impregnated.

Emerson Model 506

Continuing on the topic of oscillator circuits that use gimmicks it is found that in the Emerson Model 506 a gimmick arrangement is used to accomplish the coupling of some of the os-

FIG. 3.—Oscillator circuit of the Emerson Model 506 using gimmicks to couple the oscillator voltage to the grid of the tube.

cillator tank voltage to the oscillator grid. The circuit is shown in Fig. 3. It will be noticed that one end of coils *L1* and *L2* are open and the other ends attached, each to respective parts of the circuit. Each coil *L1* and *L2* according to our definition, represents an individual gimmick. In the preceding discussion it was shown how only one gimmick was used to formulate an effective capacity between the oscillator grid and tank circuit, but in this circuit two gimmicks are used for the same purpose and obtain a higher value of effective capacitance. The four sections of coils *L1, L2, L3,* and *L4* represent what is known generally as the oscillator-coil assembly, all these coils probably being wound on one coil form. Coil *L1* and *L4* is one continuous coil with a tapped junction to which the parallel capacitors are wired and which utilizes *L4* as the inductance of the tuned circuit. The total coil (*L1* plus *L4*) is wound in such a fashion that one end remains open and it is the end nearest the tap which forms the gimmick capacitor. Thus we have at the junction of *L4* and *C2* a separate coil *L1* existing, which has one end open and one end connected. On the same coil form is also wound coil *L2*, which is a separate coil that has one end open and one end connected to the grid of the oscillator section of the 12SA7/GT tube. Coil *L2* is wound close to coil *L1*, as shown in the diagram of Fig. 3, and between coil *L1* and coil *L2* a certain amount of capacitance exists which is determined primarily by the distance separating both coils and the common area between the surfaces of the coils. Since the coils of *L1* and *L2* are closely coupled and also since the common surface area between them is great, then a higher effective capacitance exists between both gimmicks and more oscillator voltage will be coupled to the oscillator grid.
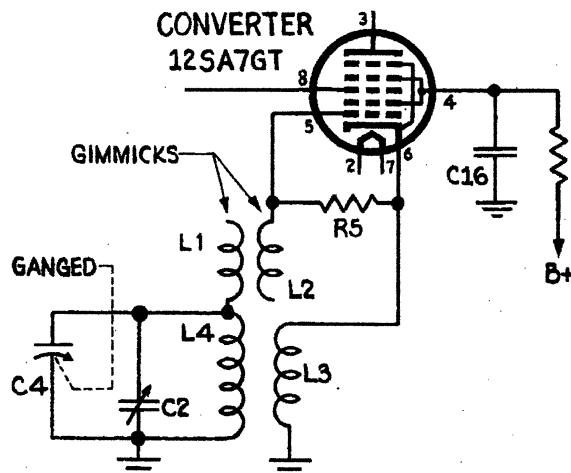
DeWald Model A602

In all the other models discussed, the gimmicks used in the oscillator circuit were employed with a tickler-coil oscillator only. In the DeWald Model A602 a gimmick is used in their oscillator circuit but a Hartley oscillator with a 12SA7 tube is employed, the circuit being shown in Fig. 4.

The gimmick coil is coupled closely to the oscillator tank coil and through the effective ca-

pacitance formed between both coils, some of the oscillator tank voltage is injected onto the oscillator grid section of the 12SA7 tube. The
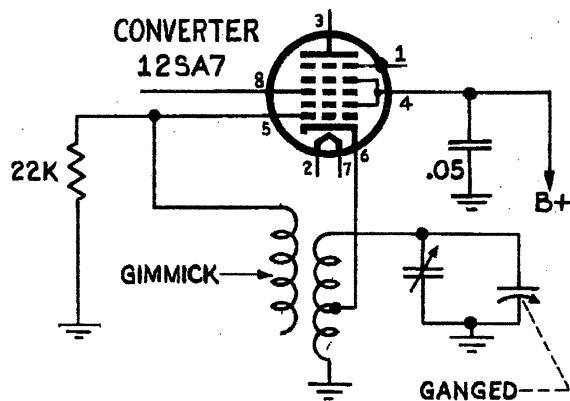

Fig. 4.—Hartley oscillator circuit used in DeWald Model A602 showing gimmick on the oscillator plate.

closeness of coupling between the coils and the amount of common area between the coils determines how much oscillator voltage will be injected onto the grid.

Espey Model XFJ-97

In the past, the usual method of coupling the oscillator tank voltage to the oscillator grid was through a fixed mica capacitor which was connected between the grid and high side of the tank circuit. It was shown in the preceding discussion how certain gimmick arrangements produced an effective capacitance and eliminated the fixed mica capacitor by coupling
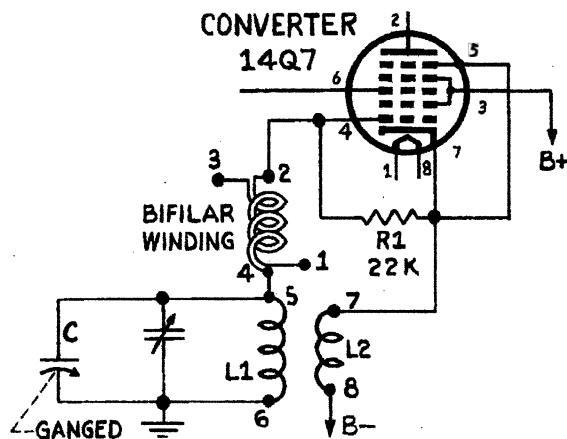

Fig. 5.—Unusual bifilar winding used as a gimmick in the Espey Model XFJ-97.

enough oscillator tank voltage on the oscillator grid to keep the oscillator functioning. This fixed mica capacitor was anywhere from 50 to 100 mmf in value. The higher the value of the capacitance, the greater would be the coupling between the oscillator tank and the oscillator grid of the tube. The previous gimmick arrangements produced a satisfactory enough capacitance, but in the Espey Model XFJ-97 a gimmick arrangement in the form of a *bifilar winding* is used to give a high capacitance. The Espey oscillator circuit is shown in Fig. 5.

Before going any further it should be understood what is meant by a bifilar winding, which was originally a method of winding non-inductive resistances. The way of forming a bifilar winding was to double the necessary wire to be used on itself; then take the loop end of the wire (i.e., that end of the wire not open) and wind the doubled wire around a form.
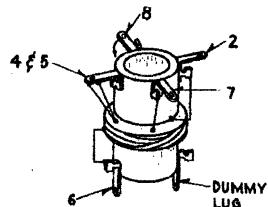
Fig. 6A. — Oscillator coil form and windings in Espey Model XFJ-97.
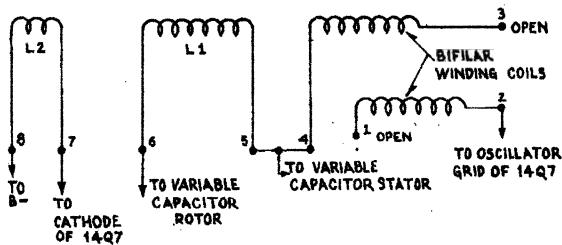
Fig. 6B.—Schematic representation of where the ends of the oscillator coils are connected.

In the circuit of Fig. 5, a bifilar type winding is used but one end is open on each side of the winding (terminals 1 and 3). There is a definite purpose for doing this and it will become evident as the complete oscillator coil is analyzed. Coils L1 and L2 represent the necessary transformer inductances incorporated with the usual tickler-coil oscillator. The bifilar winding together with L1 and L2 are all wound on one coil form. A drawing of the actual coil form with all the coils wound on it is shown in Fig. 6A and a schematic representation of the coils is shown in Fig. 6B. The

numbers from 1 to 8 on Figs. 5, 6A, and 6B are all related to one another and represent the connecting ends of the coils. The coils are wound in the following manner:

The bifilar winding starts with coil ends No. 2 and 3, the inside winding on the coil form, and the coil ends No. 1 and No. 4 are the outside terminals. Lead No. 3 is covered completely by the winding and is left open and lead No. 2 is attached to terminal 2 of the coil form. (This terminal 2 later on is conductively connected to the oscillator grid, as shown in the schematic, Fig. 5). The other ends No. 1 and No. 4, of the bifilar winding has lead No. 4 tied to terminals 4, 5 of the coil form and lead No. 1 remains open. Next coil, L1, the oscillator tank coil, is wound around the bifilar wound wires and has lead No. 5, connected to terminals 4 and 5 and lead No. 6 to terminal 6 of the coil form. The next coil L2, the tickler coil, is wound around coil L1 with leads No. 7 and No. 8 going to terminals 7 and 8 respectively. In Fig. 5, the schematic representation of the bifilar-wound wires is shown and from such a picture it is evident that the maximum amount of coupling is obtained between both gimmick sections of the windings. The actual capacitance as measured between coil terminals 2 and 3, is 125 mmf. All the coils involved are seen in Fig. 6A to be wound one on top of the other and therefore, a very high degree of coupling is obtained between the necessary windings.

International Detrola Model 571X

The application of gimmicks has found a wide variety of uses in the antenna and input circuits of many receivers. In the antenna cir-

ACTUAL INPUT CIRCUIT   EQUIVALENT CIRCUIT
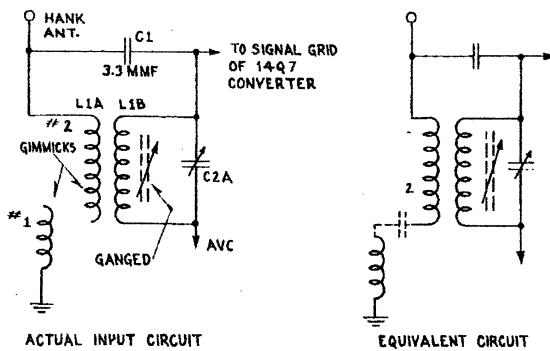
FIG. 7.—Gimmicks used in the input circuit of the International Detrola Model 571X.

cuit of the Detrola Model 571X, use is made of two gimmicks to form the necessary capacitance in completing the antenna output circuit to ground. The input circuit and its equivalent are shown in Fig. 7. The two gimmicks in the antenna circuit are wired with gimmick No. 1 having one end connected to ground with the other end open and gimmick No. 2 having one end connected to the hank antenna and the other end open. Both open ends of the gimmick lie near each other and an effective capacity exists between them, this being shown in the equivalent drawing of the input circuit by the dotted capacitor. Gimmick No. 2 is closely coupled to the permeability-tuned ganged tank circuit coil for the necessary amount of transformer coupling from the antenna circuit to the r-f circuit of the 14Q7 converter. The two gimmicks together serve the purpose of forming a complete L-C circuit from hank antenna to ground.

R-f signals are picked up by the hank antenna and the signal currents go through the gimmicks, and their effective capacity between them, to ground. The signal current passing through gimmick No. 2, which is closely coupled to r-f tank coil, sets up induced voltages into this coil L1B due to regular transformer action. The total gimmick antenna circuit can be considered as a very broad-band circuit accepting most signals. However, the tuned r-f tank circuit (consisting of L1B and C2A) when tuned to some specific station, will reflect a certain amount of impedance into the No. 2 gimmick at the particular frequency of the tuned tank circuit and there will be a maximum amount of induced signal voltage from the antenna circuit into the tuned circuit at the frequency of the station selector. The application of such gimmick arrangements saves space in the input circuit of most receivers, since by their use no fixed capacitors are needed.

Spiegel Model 433

In the Spiegel Model 433 use is also made of a gimmick arrangement in the antenna circuit but only one gimmick is employed. The antenna circuit of this receiver is shown in Fig. 8. The gimmick here is a coil of wire wrapped around the lead going from the capacity plate to the L-C r-f tuned circuit. The capacity plate serves the purpose of an indoor

antenna. The gimmick in this circuit serves the purpose of capacitively coupling the outside antenna to the L-C tank circuit for the r-f input signals. The amount of capacity is determined by how many turns the gimmick is wound around, and how closely it is wound, the lead going to the L-C circuit. Since the r-f input signals of the broadcast band cover a frequency range between 540 and 1720 kc, then the necessary capacitance needed to couple in the r-f signal need not be large and hence the gimmick need be only a few turns of wire to give the necessary capacitance coupling.
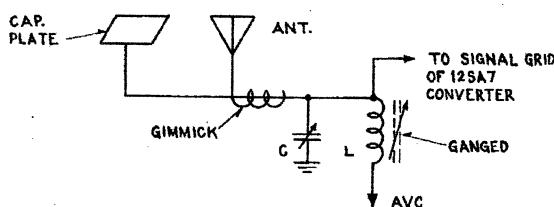

Fig. 8.—Antenna input circuit of the Spiegel Model 433 with a gimmick between the antenna and the r-f tuned circuit.

The effect of closer coupling between the gimmick and the lead wire to the L-C tank can be tested. If the gimmick is removed from the r-f lead while listening to the output of the set and is brought closer and closer to the r-f lead wire until it is as closely coupled as before, the output of the set will gradually increase. It then follows that the greater the effective capacitance between the gimmick and the lead wire to the L-C tank (i.e., the greater the coupling), the greater will be the input signal appearing on the grid of the converter tube.

Crosley Model 56FA

The Crosley Models 56FA and 56FB employ a gimmick in their antenna circuit to obtain a good frequency response of the input r-f signal. In Fig. 9, part No. 37 is the broadcast antenna coil assembly and consists of all the coils enclosed in the dashed box. All these coils are wound on one coil form with L_p and L_s representing the primary and secondary inductances respectively of the antenna input transformer for the broadcast band. The gimmick coil is attached at one end to the high side of the secondary L_s of the input transformer and

open at the other end. The open end is drawn schematically as being coupled to the primary L_p of the antenna input transformer. The primary coil 36 is from the short-wave antenna transformer and it remains in the broadcast position of the set because the switching arrangements within the set can be made more economically—and furthermore, it has negligible effect in this position.

For a moment, let us consider the broadcast-antenna input transformer No. 37 without any gimmick attached to it. The frequency response characteristic of this input transformer with its tuned secondary circuit is such that at the high-frequency end of the broadcast band, the output voltage across the secondary drops and, therefore, the level of the response curve falls off. This is an undesired feature because a frequency-response characteristic as flat as possible is required for proper reception. This unwanted feature is remedied by the insertion of the gimmick, as shown in Fig. 9. This gimmick effectively places a capacity between the high sides of the primary L_p and secondary L_s of the input antenna transformer. The usual amount of capacity in a gimmick is low and offers a low reactance at the high-fre-
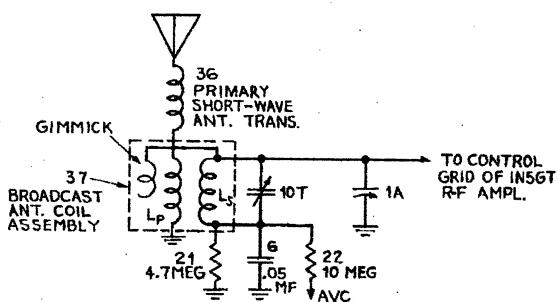
Fig. 9.—Antenna gimmick used in the Crosley Model 56FA.

quency end of the band. The way in which this capacitance, due to the gimmick, functions is as follows:

As the frequency response at the high frequencies falls off, due to the input transformer with its untuned primary and tuned secondary, the capacity formed by the gimmick compensates for this falling off by offering a low-impedance path to the high frequencies. Therefore, the gimmick allows for a better degree of coupling between the antenna and the control-grid circuit of the IN5GT r-f tube.

Firestone Model 4A24

In the Firestone Model 4A24, two gimmicks are utilized in the antenna circuit, as shown in Fig. 10, to increase the input response characteristics of the set at the high frequencies. By the use of these gimmicks an effective capacitance is added between the high sides of the primary, L_p, and secondary L_s, of the antenna input transformer. The reason why such an effective capacity is needed is exactly the same as in the preceding discussion. In this
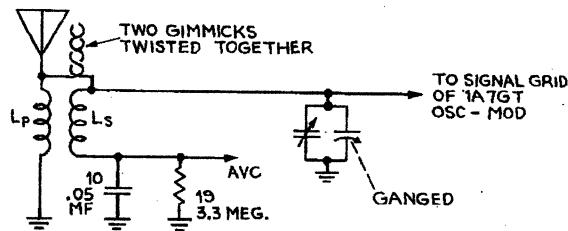
FIG. 10.—Twisted-pair gimmick used in the Firestone Model 4A24.

receiver, however, two gimmicks are used. Attached to the high side of both the primary and secondary of the input transformer is a piece of insulated wire each with one end open. Both these leads are twisted together to form capacitive coupling between them. With such a system two wires can be soldered in place readily and twisted together very quickly to give the necessary coupling for good high-frequency response.

This is a quick method to improve the response of the input system of certain receivers by the above method. In other words, if the receiver has a poor high-frequency response at the r-f input section of the set, the serviceman can solder two pieces of insulated wire to the high sides of both the primary and secondary

of the input transformer and twist them together. In that way the high-frequency response will be improved because the effective capacitive coupling between the attached wires (i.e. the gimmicks) will offer a lower impedance path to the high frequencies than will the input transformer.

Packard Bell Model 651

In the broadcast position of the Packard Bell Model 651, two gimmicks in the form of a twisted pair are used to give an effective capacitance across the secondary of the loop. The schematic representation and the actual circuit is shown in Fig. 11. The so-called twisted pair is shown schematically in Fig. 11A to be a trimmer capacitor C1. This C1 actually consists of two separate twisted gimmicks. One gimmick is attached to the high side of the secondary of the loop, as shown in Fig. 11B, and the other gimmick has one end attached to the low side of the same loop secondary, which goes to the avc circuit. Both the other ends of the gimmicks are open and both wires composing the gimmicks, are twisted together as stated above. The capacitance formed by these two twisted gimmicks is about 2 or 3 mmf. It is shown as a trimmer because by varying the degree of twist, the capacitance between both gimmicks can be changed. If the alignment procedure of this model in Vol. XV be examined, it will be noticed that no provision is made to vary the twisting of this pair of gimmicks. It appears that the manufacturer has previously set these gimmicks to give enough capacitance between them for proper tracking at the high-frequency end of the broadcast band.

Montgomery Ward Model 54 WG-2700A

In the Montgomery Ward Model 54WG-2700A, a gimmick across a capacitor is used to increase the capacity of the circuit, the schematic being shown in Fig. 12. The gimmick is tied effectively across the 4.7-mmf capacitor C32, which is connected between the 6J5 oscillator plate and 6SJ7 first detector signal-grid circuit. The 4.7-mmf capacity of C32 is not high enough to accomplish the desired coupling between the oscillator tube and the first detector tube; therefore, a gimmick is used to increase effectively the capacity of C32.

FIG. 11A.—Loop gimmick used in the Packard Bell Model 651 as shown in schematic.
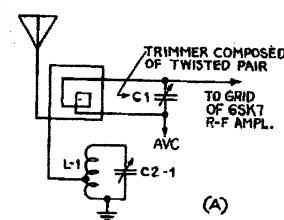
FIG. 11B.—Schematic representation of the loop gimmick.
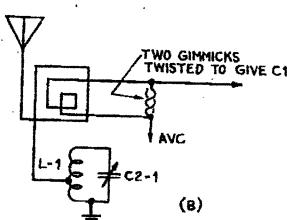
The soldered end of the gimmick is tied to that side of *C32* that is connected to the plate of the oscillator tube and the open end of the gimmick is twisted around the other side of the capacitor *C32*. A capacitance exists between the twisted wire of the gimmick and
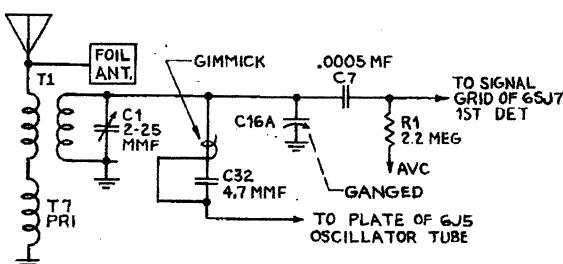

Fig. 12.—Coupling gimmick used in the Montgomery Ward Model 54WG-2700A.

the wire around which it is wrapped, which is considered to be effectively in parallel with *C32*. Since capacitors in parallel are additive then the total capacitance is increased by the exact amount of capacitance offered by the gimmick. This capacitance offered by the gimmick is low and is only of the order of a few micro-microfarads.

Galvin Model 65L11

In the Galvin Model 65L11 and 65L12, a gimmick arrangement in the r-f transformer assembly is incorporated to increase the response at the high-frequency end of the band. In this circuit, shown in Fig. 13, the r-f transformer assembly *L2* is enclosed in a shielded can and the gimmicks are considered as a part of the transformer network. Without these gimmicks the r-f transformer, with its untuned primary L_p, and its tuned secondary L_s, has a frequency-response characteristic such that at the high frequencies the response drops. Actually, the r-f transformer coupling between the 1N5GT tube and the 1A7GT tube, without the gimmicks, offers a high impedance to the high frequencies and consequently, a drop in gain at these frequencies results. The gimmicks incorporated in the *L-2* assembly effectively increase this undesired drop in response at the high frequencies. One gimmick has its soldered end connected to the high side of L_p and its open end is twisted around the open end of the other gimmick, which has its

soldered end connected to the high side of L_s. By this gimmick arrangement a capacitance is inserted effectively between the high sides of L_p and L_s. This capacitance is effective at the high frequencies since the capacitive reactance of the gimmick connection decreases with increase in frequency and results in a minimum amount of opposition to high frequencies. In other words, the impedance presented by the transformer coupling between L_p and L_s offers a greater amount of opposition to the high frequencies and therefore, the low-impedance path of the capacitance, due to the gimmick,
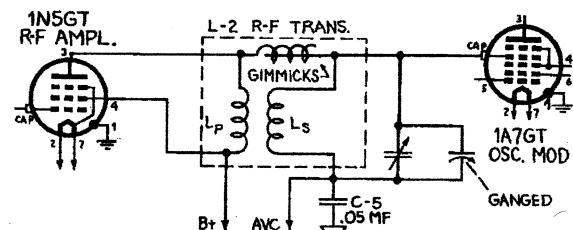
Fig. 13.—Gimmick used in the Galvin Model 65L11 to increase response at high frequencies.

compensates for the drop in response at high frequencies.

Farnsworth Model CT-060

In the Farnsworth Model CT-060, two gimmicks in the signal-grid circuit of the 12SA7 converter tube in the broadcast band are used to serve the purpose of an i-f wave trap. The circuit is shown in Fig. 14A, coils *L1* and *L2* are the two gimmicks comprising the wave trap of this set. One end of coil *L1* is attached to the signal-grid circuit of the 12SA7 tube and one end of coil *L2* is attached to one side of an .05-mf capacitor. (This capacitor is part of the avc filter network). Both the open ends of the gimmicks are facing each other and between these open ends a certain amount of capacity exists. For simplicity's sake, we will
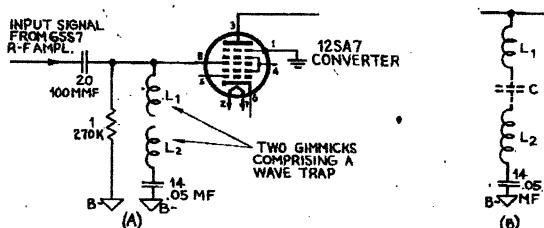
Fig. 14.—Gimmick wave trap used in the Farnsworth Model CT-060.

call this capacity C and the wave trap in an equivalent circuit is illustrated in Fig. 14B.

The capacitor C is inserted effectively between both gimmicks and its value depends upon how close the gimmicks are wound with respect to each other. Coil $L1$, the effective capacitance C, and coil $L2$ are all in series to complete an a-c signal path to the B-minus circuit for frequencies in the i-f region. In effect this $L1$-C-$L2$ circuit is a fixed series-resonant circuit for frequencies in the i-f region. This means that the inductance values of both gimmicks and the closeness of coupling between them, which determines the capacitance of C,

are the factors used in determining the resonant frequency of the wave trap.

One point of interest to the serviceman is that in doing resistance measurements and continuity checks with this model, he should remember that there are two gimmicks comprising this wave trap and that there is *no continuity through it*. It should also be understood that the series-resonant frequency of this wave trap is fixed and that there are no provisions for aligning the intermediate frequency of this trap as there usually are no other models that employ i-f wave traps in their r-f sections.

FILTERING ACTION

A filter is an electric circuit or network designed to eliminate or suppress a certain unwanted frequency or band of frequencies, at the same time selecting or allowing to pass unhampered the desired frequency or band of frequencies. From this definition, it can be seen how any tuned circuit acts as a filter, by selecting only that frequency or frequencies to which it is tuned. The intermediate-frequency (i-f) transformer of any receiver consists of parallel tuned circuits resonant to the i.f. These parallel resonant circuits offer a high impedance to the frequency to which they are tuned, (for a more thorough discussion of parallel and series circuits, see section on Coupling) thereby developing maximum currents at that frequency; whereas to frequencies that are not close to the i.f., they offer a low-impedance path in the parallel resonant circuit, thus causing decreasing circulatory currents and thereby effectively suppressing the unwanted frequencies.

Crystals also fall into the category of tuned-circuit filters, and usually are found in communication or "ham" equipment. The equivalent
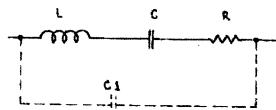
Fig. 1.—Equivalent circuit of a quartz crystal and holder.

electrical circuit of a quartz crystal is shown in Fig. 1, where L, C, and R are the equivalent inductance, capacitance, and resistance of the crystal. In order to make use of the crystal it must be inserted in a holder so that electrical connections can be made to it. This holder has the effect of adding capacity ($C1$, shown in dotted lines in Fig. 1) in parallel with the series tuned circuit. The frequency to which a crystal is resonant depends primarily upon how the crystal is cut, but it also depends upon how it is used in a circuit; that is, it can be made to respond to either its parallel or series-resonant frequency.

A quartz crystal has a very high Q, (the Q is a figure of merit or selectivity of a circuit, as explained in Coupled Circuits) and is the ratio of the inductive reactance to the resistance of a resonant circuit. Fig. 2 illustrates a crystal-filter circuit patented by Hammarlund Mfg. Co.,

Inc. and is shown as it exists in that company's Super-Pro Model 400 SX.

The crystal, $Y1$, is cut to an i.f. of 465 kc and acts as a very high Q series-resonant circuit. The added capacity of the crystal holder is neutralized by the phasing capacitor $C46$. This series-resonant circuit offers a very low impedance to 465 kc and a high impedance to other frequencies. The crystal circuit and its load, which is the parallel-resonant circuit of $L27$ and $C48$, both of them being resonant to the i.f., is actually a voltage-divider network, as shown in Fig. 3.

A 465-kc signal (resonant frequency) at the input will see a low-impedance path through $Y1$ and therefore a small amount of voltage will be developed across the crystal circuit, but this signal will then see a high-impedance load, across which the major portion of this signal voltage will be developed. If any other frequency exists at the input, it will see an increased impedance in $Y1$, therefore, a greater voltage across $Y1$ is developed. This unwanted signal then sees a decreased impedance in the parallel load circuit and will develop less voltage across the load output. This decrease in the output voltage causes the attenuation of fre-


Fig. 2.—The crystal filter circuit used in the Hammarlund Super-Pro Model 400 SX, which uses the phasing capacitor $C46$ to neutralize the added capacity of the crystal holder.

Courtesy of Hammarlund Mfg. Co. Inc.
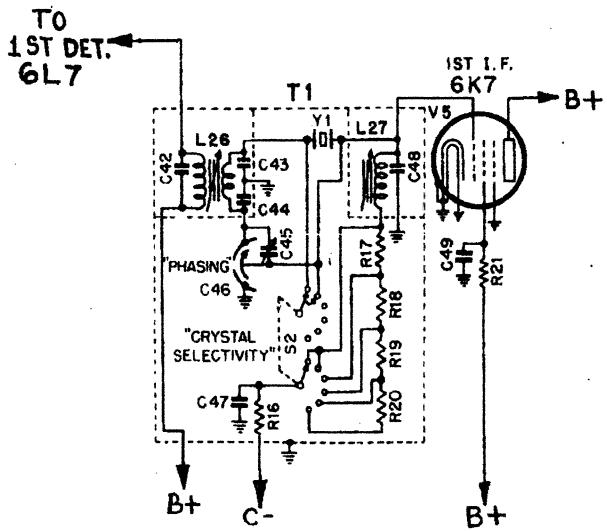
quencies other than the intermediate frequency.

This crystal filter circuit can be made more selective in steps by turning switch *S2* in Fig. 3 from position 1 to position 5, at which point it has maximum selectivity. This switch inserts additional resistance in series with *L27*, thereby lowering the Q and the impedance of the parallel-resonant load.
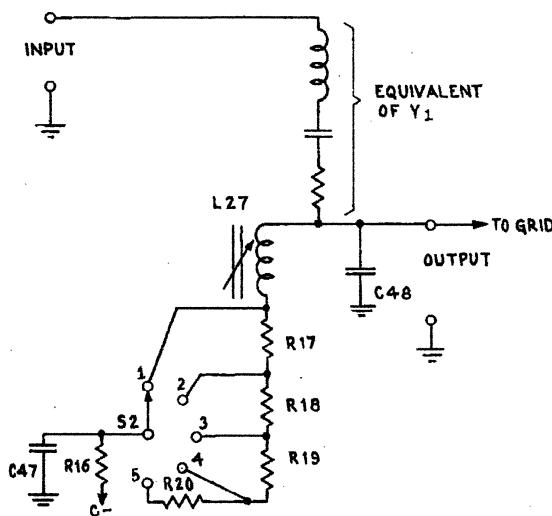

Fig. 3.—Equivalent voltage-divider network of the crystal circuit and load of Fig. 2.

The impedance (Z_P) of this load, as shown in Fig. 4, is in series with the crystal, and the Q of this crystal circuit is the ratio of the inductive reactance to the total resistance $(R+Z_P)$ that is, $\dfrac{X_L}{R+Z_p}$. In this particular circuit X_L and R have remained constant at the resonant frequency, but Z_P has decreased; therefore the ratio, which is the Q of the crystal circuit, has increased, which in turn, makes the circuit more


Fig. 4.—The impedance of the parallel-resonant load is in series with the crystal of Fig. 2.

selective. This fact will be shown more rigorously in the following mathematical analysis:
Let Q_T be the Q of the total series-resonant circuit of Fig. 4

X_L the inductive reactance of the crystal

X_{LP} the inductive reactance of the parallel-resonant circuit

R the resistance of the crystal

R_P the resistance in the parallel-resonant circuit

and Z_P the impedance of the parallel-resonant circuit, which is the equivalent series resistance added to the crystal.

$$Q_T = \frac{X_L}{R+Z_P} \text{ but } Z_P = \frac{X_{LP}^2}{R_P} \text{ therefore}$$

$$Q_T = \frac{X_L}{R+\dfrac{X_{LP}^2}{R_P}} = \frac{X_L}{\dfrac{RR_P+X_{LP}^2}{R_P}} = \frac{X_L R_P}{RR_P+X_{LP}^2}$$

$$X_L R_P = Q_T RR_P + Q_T X_{LP}^2$$
$$\text{or } X_L R_P - Q_T RR_P = Q_T X_{LP}^2$$
$$R_P (X_L - Q_T R) = Q_T X_{LP}^2$$
$$R_P = \frac{Q_T X_{LP}^2}{X_L - Q_T R}$$

Now if R_P increases then $\dfrac{Q_T X_{LP}^2}{X_L - Q_T R}$ must increase, but at the resonant frequency, R, X_{LP} and X_L remain constant, therefore Q_T is the only variable and the two possibilities are that Q_T either decreases or increases. If Q_T decreases, then the numerator $Q_T X_{LP}^2$ decreases and the denominator $(X_L - Q_T R)$ must increase, thus making the fraction, which is equal to R_P, decrease; but we know that R_P must increase, therefore Q_T cannot decrease. If Q_T increases then the numerator would increase, the denominator would decrease, thus the fraction would increase. This satisfies the condition that R_P is increased and shows that if resistance is added in series to the parallel-resonant circuit, that the total series-resonant circuit becomes more selective.

Magnavox Model CR-187

In the Magnavox Model CR-187, a parallel-resonant circuit, consisting of coil *3* and capacitor *7* in Fig. 5, is being used as a 10-kc filter between the plate circuit of the 6SN7 inverter stage and the grid circuit of the 6V6GT output stage. This filter, being parallel resonant to 10 kc, offers a very high impedance to this frequency and therefore does not allow it to pass on to the grid of the output tube, thus eliminating a 10,000-cycle whistle, which might otherwise be heard through the speaker when the set is tuned between two adjacent stations on the broadcast band.

Let us see why and how a 10-kc signal might get into the audio section of the receiver. First

of all, the Federal Communications Commission has stated that adjacent channels (stations) on the broadcast band must be 10 kc apart. Now if the set is tuned between two adjacent stations of about equal strength, let us
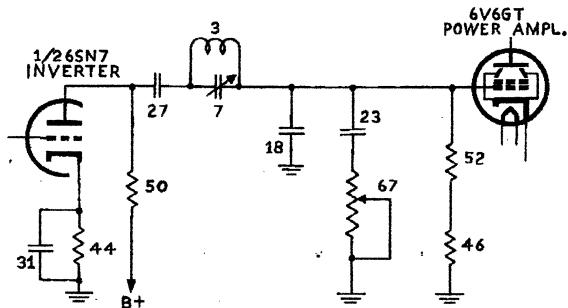

FIG. 5.—The Magnavox Model CR-187, which uses a parallel-resonant circuit consisting of coil *3* and the variable capacitor *7*, as a 10-kc filter.

say these stations are at 600 kc and 610 kc, then the set would be tuned to 605 kc. Since the oscillator tracks above the signal frequency by 455 kc, (the i.f.) to which the set is tuned, that would make the oscillator frequency 1060 kc (605+455). The incoming signals beating with the oscillator would produce signals at 460 kc (1060—600) and 450 kc (1060—610). These signals are close enough to the intermediate frequency of 455 kc to pass through the i-f transformers, and beating together, they would produce a 10-kc difference signal (460—450) which would appear as an audible whistle if not filtered out.

Emerson Model 506

In Fig. 6 is shown another type of tuned-circuit filter between the plate circuit of the 6SS7, r-f amplifier, and the grid circuit of the 12SA7GT converter tubes in the Emerson Model 506. This filter, consisting of *T1* is tuned to
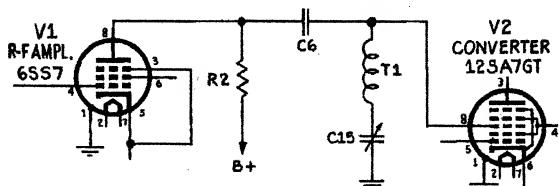

FIG. 6.—The Emerson Model 506, which uses a tuned-circuit filter consisting of *T1*, which is tuned to the i.f. by means of capacitor *C15*.

the intermediate frequency of 455 kc by means of the trimmer capacitor *C15*. This filter cir-

cuit is series resonant to the i.f. and, therefore, offers a low impedance to this frequency.

The output of the r-f amplifier consists of an untuned stage which means that it will have maximum gain at the low-frequency end of the dial; but since the filter is across the output of the r-f tube, the impedance into which the tube is working near the i.f., is low and, therefore, the gain of tube decreases at these frequencies. If a c-w signal near the intermediate frequency should find its way to the antenna, instead of being amplified and causing interference in the receiver, it will be filtered out due to the low impedance of the i-f wave trap.

The possibilities of r-f and a-f signals feeding from one stage in a receiver to another is very great, especially when the various stages are returned to a common bus, or connected through a common impedance. If nothing is done to prevent these signals from being fed back to other stages, then the varied results that might be obtained would certainly be distasteful if not harmful. These signals, depending upon how and to what stages they might be fed, can very easily cause regeneration and thus unwanted oscillations, or degeneration and distortion in the output of the receiver, or else they might also cause different tones of whistles and hum. In order to prevent these occurrences, the unwanted signals must be filtered out and not be allowed to feed back to any point where they would result in any of these harmful effects.

Olympic Model 6-501 W-V

The B— return in an ac-dc set, such as in the Olympic Model 6-501 W-V, is a common return for many of the receiver components such as

FIG. 7.—A series-resonant circuit which is used to provide a low-impedance path from B- to ground for unwanted signals.
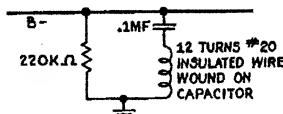
bypass capacitors, the oscillator coil, and also the cathode and grid circuits of the tubes. In order to prevent any r.f. from getting into the audio stages, a low-impedance path from B— to chassis ground is provided for these unwanted signals.

Fig. 7 shows that the low-impedance path in this model from B— to ground is a series-resonant circuit consisting of a fixed .1-mf ca-

pacitor on which is wound 12 turns of #20 insulated wire. A 220,000-ohm resistor shunts this resonant circuit, which is very broadly tuned, and therefore, any frequency above about 10,000 cycles will have a low-impedance path from the B— return to chassis ground and will not interfere with the audio output.

Espey Model 581

A common point of coupling between the grids of some of the tubes in a receiver is through the avc bus. Fig. 8 shows how the grids of the r-f amplifier, the converter, and the i-f amplifier are returned to the avc bus in the Espey Model 581.

The avc voltage is used as bias for these tubes in order to vary their gain with the incoming signal so that the receiver output will remain constant. The bias for these tubes should naturally be pure d.c., therefore, any r.f. or a.c. that might be picked up on the avc bus from one stage must be prevented from being fed back to another stage. In order to accomplish this, the grid circuits of V1, V2, and V3 must be isolated or decoupled from one another with respect to radio and audio frequencies, therefore, an R-C filter, which is called a "decoupling network" in this case, was inserted in the avc path to each grid. The decoupling filter in the grid circuit of V1 is composed of the capacitor C2 and the resistor R3. C2 offers a very low impedance to any a.f. or r.f., whereas R3 will offer a very high impedance to these frequencies, but the action of C2 and R3 to d.c.

will be just the reverse thus allowing the d.c. to pass to the grids but filtering out the audio and radio frequencies. The decoupling networks of V2 and V3, consisting of C8, R4, and R8, C19, will also act in a similar manner.

Coupling due to the common impedance of the B+supply voltage to the plates is another source of trouble. The audio frequency from the output might be fed back to a previous audio stage through the power supply in such a manner as to cause regeneration or degeneration. If regeneration occurs and is of sufficient strength, then oscillations might take place causing whistles or hum in the output and if degeneration occurs, then it is possible to have cancellation of hum but also loss of output. If a receiver is not designed specifically to have this feedback, then a filter must be used to eliminate it.

Howard Model 901-A

In the Howard Model 901-A, a decoupling filter consisting of a 47,000-ohm resistor and a .05-mf capacitor is placed in the plate circuit of the 12SQ7, detector and first-audio tube, as shown in Fig. 9. This R-C network is being used in this model as a hum filter, preventing any audio from being fed back from the 50L6 output tube through the B+ supply to the plate of the 12SQ7 and thus to the grid of the 50L6. The R-C hum filter can be considered as a voltage divider for the audio frequencies and since the 47,000-ohm resistor will offer a high impedance and the .05-mf capacitor a low impedance to these frequencies, then only a very
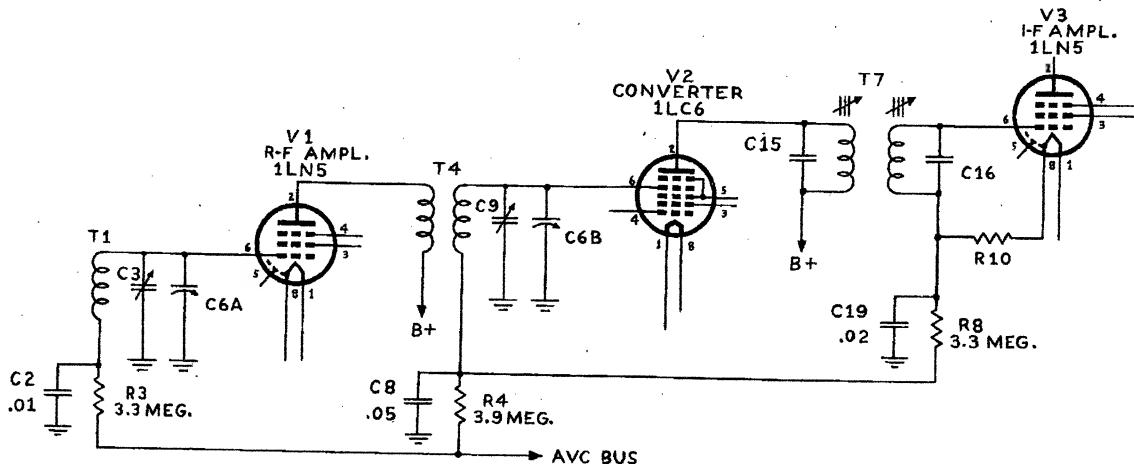

Fig. 8.—Decoupling circuit used in the Espey Model 581 showing the grids of the r-f amplifier, the converter, and the i-f amplifier returned to the avc bus.

small part of the audio will appear across the low-impedance capacitor and thus it will short out the audio frequencies effectively.

Another filter, the arrangement of which is very interesting, is also shown in Fig. 9. It is
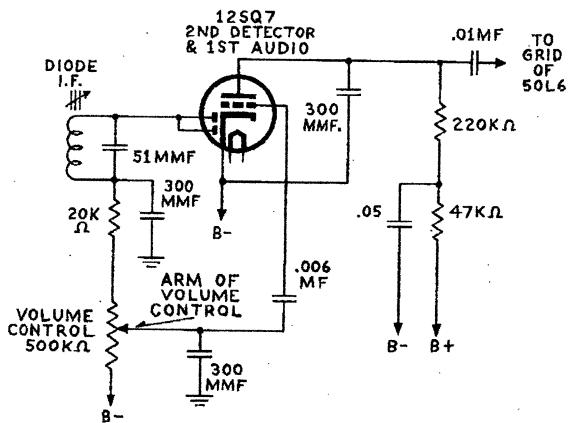
FIG. 9.—The decoupling i-f filter used in the Howard Model 901-A to filter hum; and the i-f rejection filter used in the diode second detector circuit.

the i-f rejection filter in the diode, second-detector circuit. This filter is composed of the 300-mmf capacitor, connected from the low side of the diode i.-f. transformer to ground, the 20,000-ohm resistor in series with that portion of the volume control to the point where the arm is at any particular setting, and the 300-mmf capacitor from the arm of the 500,000-ohm volume control to ground. The intermediate-frequency signal which is coupled into the diode i-f transformer, must be rectified by the diode so that only the modulation on the carrier

signal, that is, the audio frequencies, will be coupled into the grid of the 12SQ7, audio amplifier and any intermediate frequency that might appear at the diode load must be filtered out. The resistors in the i-f filter will offer a very high impedance to any i.f. in comparison to the low impedance offered by the 300-mmf capacitors. This means that there will be only a negligible amount of i.f. across the 300-mmf capacitor to ground and since the grid of the audio section of the 12SQ7 obtains its input through the .006-mf coupling capacitor across the 300-mmf capacitor to ground, then the amount of i.f. that might find its way into the grid will be negligible. This same 300-mmf capacitor, however, will offer a high impedance to audio frequencies, thereby developing a great amount of the audio across it, which is then coupled into the grid.

It is more common to see a filter of this type having both capacitors in a fixed position; that is, the capacitor which is now on the arm of the volume control, would be fixed at the junction of the volume control and the resistor in series with it. In the Howard model, the capacitor would be in such a position only when the volume is on maximum; then as the volume is decreased, the resistance between the capacitors is increased. This means that the ratio of the capacitive reactance at the intermediate frequency, to the resistance will decrease and so less of the i.f. will appear across the 300-mmf capacitor. Therefore, as the volume is being decreased, better filtering action will be taking place.

RADIO-FREQUENCY COUPLING

One of the fundamental functions in radio is the process of associating one circuit with another. This correlation is accomplished by "coupling," which is defined as *a series of methods of associating electric circuits, so that energy can be transferred from one circuit to another*. Some of the more common coupling methods are transformer coupling, capacitive coupling, resistance-capacitance coupling, inductive coupling, and many others.

By the above definition of coupling one can readily see the wide field such a topic can cover. The purpose of this section is to present an explanation of many of the different types of interesting coupling methods used in the radio receivers found in Rider's Manual Volume XV. The primary purpose is to show *why* a particular coupling method is used and *how* the transfer of energy occurs. By the relative merits of each individual type of coupling, that will be discussed, it will be possible to compare one type of coupling with another. There are a wide variety of subjects that are related to the discussion of coupling circuits that will need to be studied in order to appreciate fully the particular coupling under consideration.

Let us consider coupling methods for a moment. R-f transformer coupling, which is from the antenna input stage to the first r-f stage in which the primary of the transformer is untuned and the secondary is tuned, r-f transformer coupling that has both the primary and secondary tuned between two r-f stages, or single-tuned coupling between r-f stages as well as the simple R-C coupling usually found in many receivers, may be either series or parallel tuned. These are just a *few* of the many different types of couplings which may be employed in modern receiver design. Therefore, it will be quite necessary to study such related topics as series and parallel resonance, the Q or "figure of merit" of coils and tuned circuits, selectivity, mutual inductance, etc., before going into the analysis coupling circuits in today's receivers. The circuits chosen from the multitude of radio-service data appearing in Rider's Manual Volume XV were selected for a number of various reasons, the primary reason, of course, being that they represent new and un-

usual coupling methods. Others were chosen because it was felt that the radio serviceman should be acquainted with the functions of certain arrangements that have not been fully explained to them in the past.

Series Resonance

A typical series-resonant circuit is shown in Fig. 1. The primary elements of this circuit that determine the resonant frequency are the inductance L and capacity C. The resistance R is also introduced because invariably it is
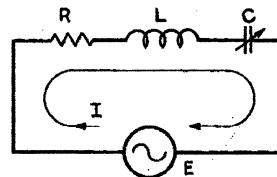
Fig. 1.—A typical series-resonant circuit wherein the I has only a single path to follow.

present in all circuits, specifically, it is the resistance of the inductance L (although in some instances a separate resistor is purposely inserted) and because in the final analysis it plays an important role with respect to coupling effects in the operation of resonant circuits. These three elements R, L, and C are connected in series and the series circuit is recognized by tracing the path of the current through the system. A circuit is said to be a series circuit if the current I is equal in all parts of the circuit, and if but a single path is offered to the flow of current, which is the case in Fig. 1, in which a generator E is assumed to be the source of the voltage applied to the circuit.

The variation of inductive reactance with frequency is shown in Fig. 2 and the way capacitive reactance varies with frequency is shown in Fig. 3. Thinking in terms of polarity and/or phase relationships of either reactance, it can then be stated, as seen from Figs. 2 and 3, that as the frequency increases, the inductive reactance increases linearly in a positive direction and also that as the frequency increases, the capacitive reactance decreases—but decreases from a negative maximum and not linearly.

Now let us once more examine the series cir-

cuit of Fig. 1. Let us suppose that the frequency of the applied voltage from the generator is increased starting from a zero frequency. What happens in this circuit? When the frequency is very low, there will be little or no
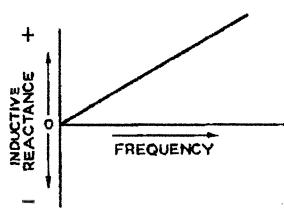

Fig. 2.—The straight graph shows that when the frequency of the current flowing through an inductance increases, the inductive reactance increases. Inductive reactance usually is considered as being positive and so is shown above the zero axis.

current flowing through the circuit; although the reactance of the coil is very small, the reactance of the capacitor at these low frequencies is very great. Now if the frequency is increased, the inductive reactance will increase from its low value and the capacitive reactance will decrease from its high value. This is in accordance with Fig. 2, wherein it is shown that the inductive reactance increases in a positive direction and in Fig. 3 that the capacitive reactance decreases numerically but always remains negative regardless of the frequency.

If one reactance is increasing and the other reactance is decreasing as the frequency is being increased, a point will be reached when the frequency will be such that the inductive reactance of the coil will exactly equal the capacitive reactance of the capacitor. Now, since the inductive reactance is positive and the capacitive reactance is negative, the two will tend to neutralize or oppose each other. Since the two reactances are equal and of opposite sign, in effect, they cancel each other and the net reactance becomes zero. This condition is termed the "resonant frequency" of L and C and it can be stated that the resonant frequency of a series circuit is that frequency at which the combined reactance of L and C in the circuit is zero.

No doubt it has been noted that in all of the considerations concerning inductive reactance and capacitive reactance, reference to the resistor R has been omitted. This is intentional since this unit, even if it represents the resistance of the coil, plays no part in the reactance phenomenon. In other words, it has a negligible effect in determining the resonant frequency of the series circuit. With zero reactance, (at resonance) the current-limiting agent is the resistance of the circuit and/or the

resistance of the generator. It is the presence of some resistance in the circuit which prevents infinite current at resonance. The *impedance*, or the opposition to the flow of alternating current, of the series circuit at resonance is its resistance. In other words, at resonance the current in the series circuit is determined by the total series resistance R in the circuit and E, the impressed voltage. Expressed mathematically, the current I in the series circuit at resonance is:

$$I = E/R$$

Off resonance, the total impedance in the circuit comprises both the resistive and reactive elements. If the resistance of the circuit is small compared to the inductive and capacitive reactance, the voltages across the reactances will exceed the impressed voltage greatly. The voltages appearing across the reactive elements of the circuit are distinguished from the *impressed* voltage by calling them *reactive voltages*. For instance, if the impressed voltage across a series-resonant circuit is 1 volt and the series resistance of the circuit at the resonant frequency is 10 ohms, then the total amount of current flowing in the circuit at resonance will be 100 ma from $I = E/R$. If the reactance of the inductance and capacitance is equal to 500 ohms
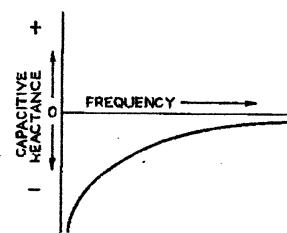
Fig. 3.—The curve below the zero axis shows that when the frequency of the current flowing through a capacitor increases, the capacitive reactance decreases. This curve is shown below the zero axis because capacitive reactance always is considered to be negative.

each at resonance, then the reactive voltage drop across either L or C will be the current flowing in the circuit (100 ma or 0.1 ampere) multiplied by the reactance of the unit across which this reactive voltage is measured. Consequently, the reactive voltage drop will be 500 ohms $\times$ 0.1 or 50 volts.

The resonant frequency of this circuit is determined from the relation that at resonance, the inductive reactance equals the capacitive reactance. And since the inductive reactance is equal to $2\pi f_r L$ and the capacitive reactance is equal to $\dfrac{1}{2\pi f_r C}$, then:

$$2\pi f_r L = \frac{1}{2\pi f_r C}$$

$$(2\pi f_r L) \times (2\pi f_r C) = 1$$

$$(2\pi f_r)^2 LC = 1$$

$$f_r = \frac{1}{2\pi \sqrt{LC}}$$

where f_r equals the frequency at resonance and π a numerical value equal to 3.14.

What happens if the frequency is increased beyond the resonant frequency? Referring once more to Figs. 2 and 3, it can be seen that the reactance of the coil will increase, whereas the reactance of the capacitance will decrease. Because of this variation, the balance which existed between the two reactances at resonance is destroyed.

At resonance, the total impedance of the circuit is a minimum and is equal to the resistance in the circuit. At frequencies above and below resonance, the reactive elements have to be taken into account in obtaining the total impedance and this total impedance, above or below resonance, therefore increases. It should be remembered, however, that the impedance of the circuit, off resonance, is *not* just a pure addition of the resistance of R and the reactance of the other elements, but the addition of the components is accomplished vectorially. The analysis of this type of vectorial addition is too complex to be considered here and also would require too much detailed work.

The curve of how the impedance in a series-resonant circuit varies with frequency is illustrated in Fig. 4A, and the curve of how the current in the same circuit varies with frequency is shown in Fig. 4B. From these curves it can be concluded that at resonance, in a series-tuned circuit, *the current is at a maximum and the impedance a minimum;* also it
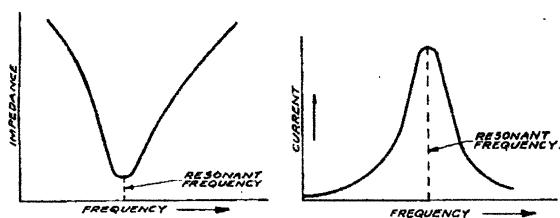
Figs. 4A, left, and 4B.—The left graph shows that the impedance is a minimum at the resonant frequency for a series circuit and the graph on the right shows that the current is a maximum at the resonant frequency.

will be noted that when the current is a minimum the impedance is a maximum. The reason why the current does not become infinite at resonance is because the circuit impedance is never zero, due to the resistance of the coil. Fig. 4B is said to be a "resonance" curve for a series-resonant circuit.

Parallel Resonance

In addition to series resonance, there is another type of resonant circuit which appears in practically all radio circuits. This is the parallel-tuned circuit and resonance in such a circuit is termed *parallel resonance.* As in the case of the series-resonant circuit, the resonant frequency of a parallel circuit is reached when the reactance of the coil is equal to the reactance of the capacitor.

An example of a parallel-resonant circuit is illustrated in Fig. 5. The reason why the re-

Fig. 5.—A typical parallel-resonant circuit wherein the line current can follow two paths, instead of one path as in the circuit of Fig. 1. The lower reactance in the parallel circuit influences the current flow.
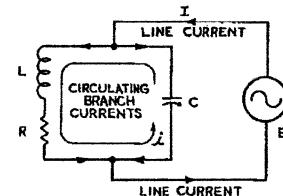
sistance is shown in the inductive branch and not in the capacitive branch is because as far as resistance is concerned, the coil is the unit that contains the greatest amount of resistance, whereas the capacitor has a negligible amount of resistance. The distinction between a series and a parallel tuned circuit is in the manner in which the current flows through the circuits. We have stated that a series circuit is one in which there is but one path for the current. In a parallel circuit, however, there is more than one path for the current. In Fig. 5 will be noted a separate capacitive branch and a separate inductive branch. The presence of R in the inductive branch will receive additional mention later.

Let us consider for a moment each of these branches separately with respect to the frequency of the applied voltage. At very low frequencies, the reactance of the capacitor branch is high (refer to Fig. 3) so very little current flows in that circuit. However, at the low frequencies the inductive reactance of L is low and, consequently, the inductive branch carries the major portion of the current. At very high fre-

quencies, the reactance of the inductive branch is very high, (refer to Fig. 2) so very little current flows through that branch; however, at such high frequencies the reactance of the capacitive branch is very low, so that the major portion of current flows through the branch. As in the case of the series-resonant circuit, at some frequency in between these two extremes, the impedances of the capacitive branch and the inductive branch are the same. This is termed the "resonant frequency" of the parallel circuit.

In the series circuit, it was shown that the limitation of current was due to the higher of the two reactances. Whether the capacitive reactance or the inductive reactance was high, as a result of the frequency of the applied voltage, the current through the circuit was small. In the parallel-resonance circuit, we find a different condition. In this circuit the dominating influence, with respect to current flow, is the *lower* reactance. If the frequency is such that the capacitive reactance is high and the inductive reactance is low, the total impedance of the circuit is low, because a low reactance shunts the high reactance. If, on the other hand, the frequency is high, so that the inductive reactance is high, but the capacitive reactance is low, the net impedance is still low for the same reason as given above. At resonance, however, the capacitive reactance and the inductive reactance are of like order and the total impedance is a maximum.

Equal current flows through the two branches, but the current through the coil is at every moment opposite in phase to the current through the capacitor. This is so because the current through the coil lags behind the voltage across it, whereas the current through the capacitor leads the voltage across it. The sum result is that the total line current through the circuit is very small, consequently, the impedance of the circuit at resonance is a maximum. In a parallel-resonant circuit, when the resistance in the circuit is small compared to the reactance of L and C, the resonant frequency is given by the same formula as in the series-

resonant case where $f_r = \dfrac{1}{2\pi\sqrt{LC}}$

The reference to the minimum current through the parallel circuit should not be confused with the fact that at resonance there is

very large circulating current *in the tuned circuit itself*—that is, flowing around L and C, as indicated in Fig. 5. In fact, this circulating current i in the tuned circuit may be many times larger than the line current I flowing through the tuned circuit. This large circulating current, which is present only at resonance, is due to the energy surging back and forth from the capacitor to the coil. Furthermore, because the capacitive reactance balances or offsets the inductive reactance, the net reactance in the tuned circuit is zero. Consequently, the tuned circuit behaves as a pure resistance and may be considered as such when computed in radio circuits. The smaller the resistance R of the tuned circuit, the greater will be the circulating current in the tuned circuit at resonance and the greater is the operating efficiency. At resonance, however, the pure resistive network that the parallel circuit offers is not equal to just the resistance of the circuit but to a rather more complicated expression. At resonance, the parallel circuit offers an impedance that is purely resistive in nature and equal to $\dfrac{(X_L)^2}{R}$ where X_L

is equal to the inductive reactance of the circuit.

It might be well to elaborate upon the relation between the circulating current in the tuned circuit and the line current through the tuned circuit. During the state of resonance, the function of the line current is to be just great enough to supply the circuit losses. In view of the low resistance of the tuned circuit, the energy losses therein are small, therefore, the line current is small. The higher the losses in the tuned circuit, the higher is the line current with respect to the circulating current.

The manner in which the impedance of a parallel circuit varies with frequency as resonance is approached and passed, is shown in Fig. 6. Compare Fig. 6 with Fig. 4A. Note that

Fig. 6.—The impedance of a parallel-resonant circuit is a maximum at the resonant frequency, which is opposite to the condition for a series-resonant circuit shown in Fig. 4A.
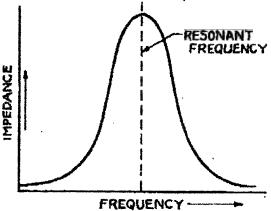
at resonance, the impedance of the series circuit is a minimum, whereas in the parallel circuit, it is a maximum.

TABLE 1

	Series-Resonant Circuit	Parallel-Resonant Circuit
Impedance Z	Minimum	Maximum
Line Current I	Maximum	Minimum
Relation between inductive and capacitive reactance	Inductive reactance equals the capacitive reactance	Inductive reactance equals the capacitive reactance
Voltage across C	Maximum due to maximum line current	Maximum due to maximum circulating current in parallel branch
Voltage across L	Maximum due to maximum line current	Maximum due to maximum circulating current in parallel branch
Comparative voltages across C and L	Equal and a maximum due to equality of reactances and same current flowing through each	Equal and a maximum due to equality of reactances and same circulating branch current

At this point it will be well to tabulate the differences between parallel- and series-resonant circuits as explained above; this is shown in Table 1.

"Q" or Figure of Merit

Since coils and tuned circuits are used to such a great extent in radio, a fundamental expression and/or letter to represent the quality or figure of merit that a coil or circuit possesses, has been chosen, the letter "Q" having been chosen for this expression. The best means of expressing the quality that a coil or circuit possesses is *the ratio between the inductive reactance and the total effective series resistance.* Sometimes a difference is made between Q's representing coils and those representing circuits by special reference to the *coil Q* or the *circuit Q.* When the Q of a coil or circuit is high, it is said to have good quality and in many cases such high Q coils and circuits are desired for good selectivity. More will be said about selectivity later on.

At radio frequencies, the resistance of a coil increases with frequency (that is why reference is made to a-c resistance as distinguished from d-c resistance) due to special factors known as *skin effect, proximity effect, dielectric losses,* and *eddy currents.* Also with an in-

crease in frequency, the reactance of a coil increases so that within a limited range of frequencies the Q of a coil or a circuit may remain substantially constant.

There are some fundamental relations about the voltage and current in series- and parallel-resonant circuits that are best expressed in terms of the circuit Q's. For instance in a series-resonant circuit, the voltage across either the L or C experiences a resonant rise in voltage which is equal to the Q of the circuit multiplied by the impressed voltage. Similarly, in a parallel-resonant circuit, the current flowing through either the coil or capacitor experiences a resonant rise in current which is equal to the Q of the circuit multiplied by the line current.

When the Q of a circuit is low, the selectivity or ability of the tuned circuit to pass a small given band of frequencies decreases and some undesired frequencies will also be passed. When the Q of a circuit is high, the selectivity of the circuit is said to be very sharp and the tuned circuit accepts only the desired frequencies. From the above analysis of what Q is and how often it is used to represent certain characteristics of coils and tuned circuits, it readily becomes evident how important this feature is.

SELECTIVITY CHARACTERISTICS

Effect of Resistance on a Series-Tuned Circuit

Due to the relation between the inductive reactance and the capacitive reactance in a series-tuned circuit, there is no net reactance at resonance; also, the current at resonance is limited solely by the resistance of the circuit. The greater the resistance, the lower will be the current. The variation of current at resonance in a series circuit for constant voltage input and for varying values of resistance is shown in Fig. 7. This curve is the equivalent of a resonance curve for such a series-resonant circuit with various values of resistance.

In referring to the resistance R, it is not necessary that this be an external resistor or an individual resistor, but that this symbol may represent the resistance of the coil, or, it may represent a poor contact in the circuit or the electrical connections in the system. The three values of resistance shown are arbitrary, selected to illustrate the effect of resistance in the circuit. Since every circuit has a definite amount of inherent resistance, so that a specific finite curve exists, it then becomes necessary, in order that the maximum current be obtained at resonance, to make the circuit structure such that it possesses the minimum amount of resistance. This is a matter of design.

Now—in addition to the fact that the current at resonance is decreased as the resistance is increased—a close examination of the curves shows that another significant fact is developed as the resistance is increased. If this current curve is representative of the ability of the circuit to discriminate between frequencies both sides of the resonant frequency, then it is evident that as the resistance is increased, not only does the current decrease, but the ability to discriminate between frequencies likewise is impaired. This reference to discrimination between frequencies, sometimes termed "selectivity" deserves lengthier comment. The ability of a circuit to discriminate between frequencies both sides of resonance depends primarily upon the ratio between the current in the circuit at resonance and the current in the circuit at some frequency either side of resonance.

For example, referring to curve No. 1 of Fig. 7, if we set an arbitrary value of 1000 kc

as being the resonant frequency, and if we say that the current in the circuit is 10 units at resonance, the characteristic of that circuit, at whatever value of resistance is present, is such that at 10 kc off resonance, the current in the system is reduced to 6 units. In other words, if the circuit is resonated to 1000 kc, and some other signal of 990 kc enters the circuit, the current at this frequency is only 6 units. The same is true if another signal of 1010 kc enters the circuit when the circuit is resonated to 1000 kc. The greater the current at the resonant frequency, with respect to the current in the circuit at some frequency off resonance, the more readily does the circuit discriminate between the two frequencies, namely, the resonant frequency and the off-resonance frequency. Referring to curve No. 2 of Fig. 7, representative of the same circuit but with a larger value of resistance, it will be seen that the circuit is still resonant to 1000 kc, but now the current at resonance is only 8 units, and at 10 kc off resonance the current is still 6 units. This means that the circuit will not discriminate as readily between the 990-kc and the 1000-kc signals.

Referring to curve No. 3 of Fig. 7, which is representative of a still greater value of resistance present in the resonant circuit, the circuit is still resonant to 1000 kc, but the curve is much flatter; it is no longer as steep as the curve developed with lower values of resistance.
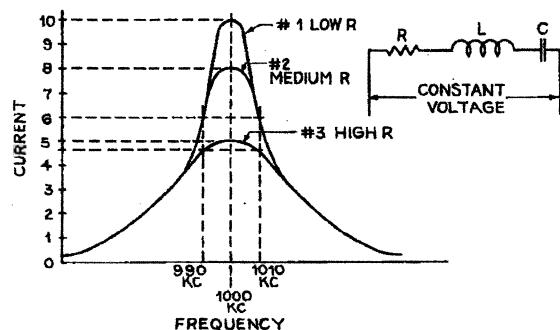
FIG. 7.—The three curves show how the current varies at resonance in a series circuit, when the resistance is changed in a circuit such as that shown above. The greatest current flow is obtained when the resistance is a minimum.

If the current at resonance is checked, it will be found that it is 5 units, and at 10 kc off resonance, it is approximately 4-3/4 units. The off-resonance signal develops approximately as much current in the system as the resonant

signal. Such a circuit would not discriminate between the two frequencies, and if such a condition is interpreted in actual practice, it means that the off-resonance signal would interfere with the resonant-frequency signal. What has been said in connection with the off-resonance signal at 10 kc below the resonant frequency, is likewise true if the interfering signal or the off-resonance signal is 10 kc above the resonant frequency.

The ability of a circuit to discriminate between frequencies is the ability of that circuit to *select* signals and is a measure of the *selectivity* of the circuit. The flatness of the high-resistance curve 3 implies directly that the circuit will not be able to respond differently to different frequencies and hence implies that the selectivity of the circuit is poor. Accordingly, it can be seen that the greater the resistance of the circuit, the less its selectivity power. Interpreting this entire situation into practice, it would mean that interference would exist and that the amplitude of the signal (height of curve) at the resonant frequency would be comparatively low.

Effect of Resistance on a Parallel Tuned Circuit

It is somewhat more difficult to understand the effect of resistance on the characteristics of the parallel tuned circuit without going into mathematical computation at the moment. However, if the explanation of a tank circuit given in the section on "Oscillators" is studied, it can be understood how, at resonance, the electrons surge back and forth from the coil to the capacitor. With a minimum amount of resistance in the circuit, this energy transfer

FIG. 8.—When the resistance in a parallel-resonant circuit is a minimum, the impedance is a maximum at the resonant frequency of the circuit, shown above. Notice how the resonance curves flatten as the resistance is increased.

can take place with maximum amplitude at the resonant frequency. Consequently, the circulating current in the *tuned* circuits is high. Adding resistance to the tuned circuit, such as adding it in series with the inductive branch, does two things: it reduces the amount of current circulating between the coil and the capacitor, in other words, it damps the circuit; and it reduces the selectivity.

In accordance with what has been said before concerning the function of the line current, loss is introduced as a result of the resistance and tends, in effect, to reduce the impedance of the parallel-resonant circuit. With a reduction in the impedance of the circuit, the voltage developed across the circuit is reduced. Because of the damping of the circuit, the change in impedance both sides of the resonant frequencies is not as great as when minimum resistance exists in the circuit. This means that the selectivity powers of the circuit are impaired as the resistance of the circuit is increased, or, if it is said that such resistance represents losses—as the losses in the circuit are increased. Fig. 8 illustrates the variation in impedance of a parallel circuit at three values of resistance over a band of frequencies.

The low-resistance curve No. 1, Fig. 8, represents what would be the equivalent of a minimum, normal amount of resistance in such a circuit. Note the sharpness of the resonance curve, which is similar to the curve No. 1 in Fig. 7. In this curve the discriminating powers of the circuit are illustrated by noting the change in impedance as the frequency of the applied voltage is made lower and higher than the resonant frequency. In actual function, Figs. 7 and 8 are similar. Both are selectivity curves, but in Fig. 7 the resonance curve is established on the basis of current change in the circuit, whereas in Fig. 8 the resonance curve is developed as a result of the change in impedance.

Examining curve No. 1 in Fig. 8, it will be noted that the impedance at resonance is 100,000 ohms and at 990 kc, which is 10 kc below resonance, the impedance is approximately 60,000 ohms. Examining curve No. 2 in Fig. 8, which is for a higher value of resistance, the impedance at resonance is approximately 80,000 ohms, and at 990 kc, the impedance is approximately 60,000 ohms. Obviously, the change in impedance as the fre-

quency is changed by 10 kc, is not as great as previously, and if the 990-kc voltage is that of an interfering signal, the selectivity powers of the circuit would not be great enough to choose between the two signals. Note also that the slope of the curve is not as steep, and that the peak of the resonance curve is much broader than the curve No. 1, indicative of the lower value of resistance. Curve No. 3 of Fig. 8 is for a still greater value of resistance. Note that the top of the curve is much less sharp. At resonance, the impedance is approximately 50,000 ohms, and at 990 kc, the impedance is approximately 48,000 ohms—practically no variation in impedance over this 10-kc band. Obviously, the selectivity of such a circuit over this 10-kc band is substantially nil. What has been said concerning the 10-kc change below the resonant frequency naturally is applicable to 10 kc above the resonant frequency.
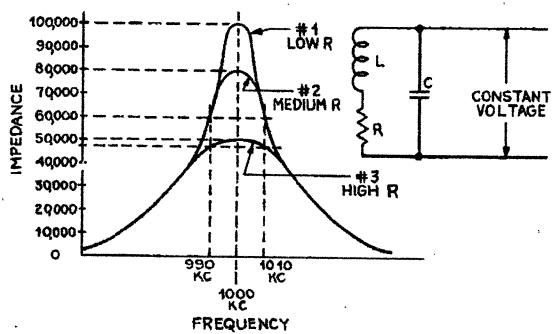
In accordance with the explanations that have been given in connection with Figs. 7 and 8, it is evident that resistance in either series- or parallel-resonant circuits can be very harmful and should be kept at that minimum dictated by requirements. At the same time, it should be mentioned that under certain conditions, a small value of resistance must be inserted into a circuit, in order to increase the width of the response curve. Such is the case in high-fidelity systems in some of our modern receivers.

The series resistance in a series- or parallel-resonant circuit has been discussed. It is equally important to keep the shunt resistance across a tuned circuit as low as possible. Effectively, a resistance shunted across a tuned circuit is equivalent to a resistance in series with the coil. Shunt resistance will interfere with the selectivity characteristics just as readily as series resistance, but sometimes a shunt resistance is desirable in the case of transformer coupling. This resistance often is inserted in a parallel-tuned circuit to broaden the response curve of the tuned circuit so that the discrimination between frequencies is not so great and a wider band of frequencies will be passed.

Self Inductance and Mutual Inductance

Even though there is no direct electrical means of connection between circuits to be transformer coupled, there is an invisible means of transferring the desired energy from one circuit to another. In "transformer action" this invisible means of transferring the energy is found and it is due to a property known as the "mutual inductance" of a transformer. At first it will be better to understand how a coil alone acts toward an alternating voltage and then the realization of mutual inductance will be understood more easily.

When an alternating voltage is applied across a coil, an alternating current flows through it. This flowing current varies between a minimum and a maximum and changes its polarity in direct accordance with the rise and fall and changes in polarity of the impressed alternating voltage. In other words, when the alternating voltage is at any one instant a positive maximum or a negative maximum, the alternating current is at that instant zero; and when the alternating current is at any one instant a positive maximum or a negative maximum, the impressed alternating voltage is passing through a zero point. Now, as this current intensity continually changes and reverses in direction, a changing magnetic field is set up around the turns of the coil itself which in turn is varying continually.

This magnetic field sets up varying magnetic lines of force (called "flux") which cut the turns of the coil. This varying magnetic force induces into the coil a voltage (known as a "counter or back electromotive force"), which in turn produces an alternating current that opposes (is in opposite phase) the original alternating current. Therefore, it is seen that as the original alternating current continues to increase, there is likewise an increasing magnetic force which cuts the turns of the coil. This magnetic force produces an increasing opposing current to the original current in the coil. Actually, a phenomenon, where an
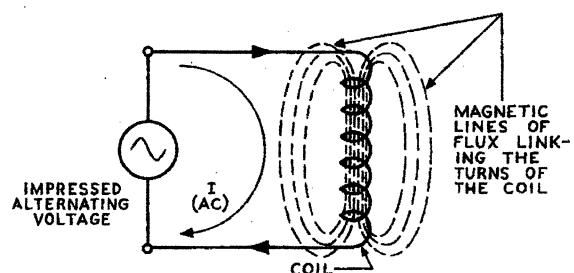

Fig. 9.—When an alternating current flows in a coil, magnetic lines of flux are produced about the coil, which interlink with the turns.

increase in the original current flowing through the coil causes an increase in the introduction of an opposing current, which tends to retard the increase of the original current, exists. When an alternating current passes through a coil as shown in Fig. 9, a magnetic flux is produced that encircles the coil and likewise the current passing through the coil. There is a relationship between these lines of flux and the amount of current flowing through this coil that gives the coil the property of inductance (sometimes known as the "coefficient of self inductance" of the coil). The unit of measurement of inductance, L, is the henry.

The relationship between the inductance of the coil, the amount of flux linkage produced, and the current producing it is given by the following mathematical expression:

$$Inductance = \frac{Amount\ of\ flux\ linkage}{Current\ producing\ flux} \times 10^{-8}$$

expressed in henrys

where the current is measured in amperes. Therefore, from the above relation, inductance can be defined as the amount of flux linkage per ampere of current producing the flux.

Now if another coil were put close to the one shown in Fig. 9, it can be seen from Fig. 10 that the flux lines from coil L_p are linked through L_s. These two coils with self inductances L_p and L_s, are said to be coupled inductively. A mutual inductance, M, therefore, exists between them. That is to say, the flux lines from coil L_p induce a voltage into coil L_s (sometimes known as the "induced electromotive force," abbreviated e.m.f.) and the amount of voltage induced is dependent upon the number of lines of flux from coil L_p encircling coil L_s. In effect then, when mutual inductance is said to exist between two coupled
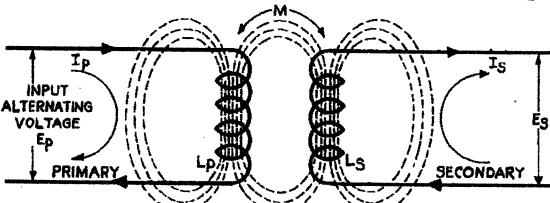

FIG. 10.—When an alternating current flows in the primary coil of a transformer, the magnetic lines of flux interlink with the turns of the secondary coil, which is coupled to the primary. This interlinkage of flux induces in the secondary coil a voltage, the magnitude of which depends on the ratio of the number of turns in the two windings and the closeness of coupling between both coils.

inductances, it actually means that there is a magnetic field common to both inductances. From the above brief analysis it is readily seen how there is an energy transfer from one circuit to another without any direct electrical means of connection between these circuits.

Similar to self-induction, the unit of measurement of mutual inductance is the henry. Mutual inductance, symbol M, can also be defined as a ratio between flux linkage and current flowing. Mathematically it is:

$$M = \frac{Flux\ linkages\ in\ second\ coil\ produced\ by\ current\ in\ first\ coil}{current\ in\ first\ coil} \times 10^{-8}$$

or

$$M = \frac{Flux\ linkages\ in\ first\ coil\ produced\ by\ current\ in\ second\ coil}{current\ in\ second\ coil} \times 10^{-8}$$

where the current is expressed in amperes. Therefore, mutual inductance is defined as the amount of flux linkage with one coil per ampere of current flowing in the other coil. A very important thing to remember is that the sign of mutual inductance (positive or negative) depends on the relation of mutual flux to self flux and the assumed positive directions of voltage and current.

The above mathematical relations for mutual inductance M are not easy to handle from the standpoint of simple mathematical computations because they involve breaking down the flux linkages into their components, and delving into the theory and explanations of flux and lines of flux, so mutual inductance is mathematically defined in another way. If the wave shape of the current flowing in the first circuit or primary circuit is sinusoidal (i.e., like a sine wave), then the voltage induced in the second circuit due to the current in the primary circuit is equal to the mutual reactance multiplied by the current flowing in the first or primary circuit. That is:

$E_s = -2\pi f\ M\ I_p$, and rearranging we have

$$2\pi f\ M = -\frac{I_p}{E_s}$$

Here $2\pi f\ M$ is a new quantity called the "mutual reactance" of the circuit and it is seen

to be a negative ratio. If the circuit of Fig. 10 were applied to this equation it would hold true, that is according to the directions of current chosen. In this instance, E_s is the induced voltage appearing across the open terminals of the secondary circuit, I_p the current flowing in the primary circuit, and $2\pi f$ is an expression of frequency relative to the rate of change of the primary current, where f is equal to the frequency in cycles per second.

The main reason for giving the negative sign to the equation has as its basic foundation a fundamental law of electricity—namely Lenz's Law. Applying Lenz's Law to the problem here, it can be stated that an induced current (or induced voltage which sets the current flowing) due to magnetic lines of force from one coil linking the other, is always in such a direction as to oppose the magnetic field change that is producing it. That is the reason why induced voltages are sometimes referred to as "back electromotive forces" (B.E.M.F.) or "counter electromotive forces" (C.E.M.F.)

Inductors, like resistors, follow the same mathematical laws when connected in series and parallel.

For a *series* circuit

$$L_{total} = L_1 + L_2 + L_3 + \text{the inductance of any and all other coils used in series connection}$$

For a *parallel* circuit

$$\frac{1}{L_{total}} = \frac{1}{L_1} + \frac{1}{L_2} + \frac{1}{L_3} + \text{the reciprocal of any and all other separate coils used in the parallel connection.}$$

There must be no coupling effect between these inductors connected in series or parallel. If there is some degree of coupling, between two inductors L_1 and L_2 connected in *series*, as shown in Figs. 11A and 11B, mutual inductance exists between them and the total inductance in series for this circuit will be

$$L_{total} = L_1\,L_2 \pm 2M$$

$+2M$ if the coils are connected in series aiding and $-2M$ if they are connected in series opposing. (See Figs. 11A and B.)

Reflected Impedance

To understand fully the different types of transformer coupling, certain fundamental properties related to such coupling arrangements should be understood; especially since transformer coupling is used to a great extent in r-f and i-f circuits. There are certain impedance relations concerning this type of coupling that are considered the underlying characteristics upon which the selection of these transformers are based. One of the most important properties of transformer coupling that is often misunderstood and rarely discussed is the topic of "reflected impedances".

Assume that an r-f transformer in operation in some particular circuit is being studied. Looking *into* the primary of this transformer, what impedance would be seen? Would it be just the impedance of the primary circuit components alone or some other impedance? If some impedance other than that of the primary circuit alone, then exactly what would it be? These are the type of questions one should ask himself if he wants to understand how such factors as the closeness of coupling, mutual inductance, and resonance affect the selectivity of the r-f transformer under consideration. Due to the existence of a mutual inductance between the primary and secondary of the r-f transformer there is reflected into the primary circuit (looking into the r-f transformer through the primary terminals) a certain amount of impedance. This reflected impedance is added to the rest of the impedance of the primary circuit to give a total *effective* primary impedance that is different from the impedance of the primary when considered alone. The amount of reflected impedance can be actually calculated and for the case under consideration, it is equal to the ratio of the mutual
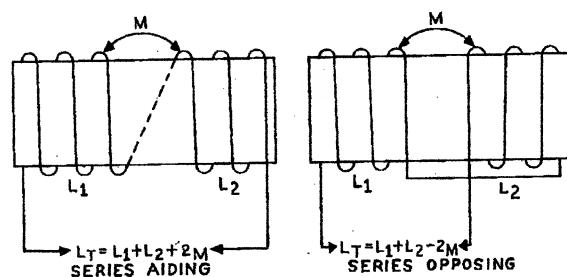
Figs. 11A, left, and 11B.—The total inductance of two coils connected in series aiding, shown on the left, is greater than when they are connected in series opposing.

reactance squared to the total series impedance of the secondary circuit. Mathematically the reflected impedance, sometimes called the "coupled impedance", is given by the following relation:

$$Reflected\ Impedance = \frac{(2\pi fM)^2}{Z_s}$$

where Z_s is equal to the series impedance of the secondary circuit alone. In other words, the secondary current, caused by the voltage induced into the secondary circuit, flows through all the secondary circuit components in a series path.

The total primary impedance of the r-f transformer under consideration is then equal to the effective series impedance of the primary circuit, considered alone, plus the reflected impedance of the secondary into the primary. It can also be shown that when looking into the r-f transformer from the secondary side, the primary circuit likewise reflects an impedance into the secondary circuit. Consequently, the total secondary impedance, when looking into the secondary terminals, is equal to the effective series impedance of the secondary circuit plus the reflected impedance of the primary circuit.

RESISTANCE—CAPACITANCE COUPLING

Perhaps the most important type of coupling where there is a direct physical connection between both circuits to be coupled is "resistance-capacitance" coupling. Between transformer coupling and resistance-capacitance coupling, the field of coupling is estimated to be over ninety percent covered; all other types of couplings are modifications of the above or are complex couplings involving each type separately or both types together. "Resistance-Capacitance" coupling (usually abbreviated as "R-C coupling") is sometimes referred to as just "resistance" coupling.

In R-C coupling, three important circuit elements must be taken into account. In Fig. 12A is shown a typical circuit of a resistance-capacitance coupled amplifier in which the three important elements are the plate load resistance R_b, the coupling capacitor, C_b (sometimes known as a "blocking" capacitor, as will be seen later), and the grid-leak resistance R_c. The reasons and uses for each of these will be taken up separately to explain how each operates individually and how necessary it is to have them all operating in unison.

Assuming the capacitor C_b is the only element connected in the circuit, as shown in Fig. 12B, and that the proper potentials are applied, then plate current will flow. This plate current means a stream of electrons flows from tube V_1 in the direction of the arrows shown in Fig. 12B and it will break up into different paths, some going to the positive side of the B supply and the others impinging against the left-hand plate of the capacitor C_b. The electrons accumulating on the capacitor plate, place a negative charge on this plate and thereby cause electrons to be driven away from the right-hand side of the capacitor to neutralize the potentials across C_b. Since the right-hand side of the capacitor was originally neutral and a negative charge is removed from this neutral side, then a positive charge must remain. These electrons flowing from the capacitor to tube V_2, accumulate on the grid of V_2. If the original plate current of tube V_1 is fluctuating due to the alternating input signal E_1, then the electron stream flowing to tube V_2 will likewise be fluctuating and this will in turn cause a pulsating or fluctuating current to flow in tube V_2.

So far the operation of a straight-forward system of capacitor coupling alone has been considered. The main disadvantage of using C_b alone is that very little or no signal current will flow through it, most of the current going to the B supply. Considering the electron or current distribution on the basis of two impedances in parallel (i.e., the reactance of the capacitor C_b in series with the reactance of the grid-to-cathode interelectrode capacitance C_{gk} as one impedance in parallel with the internal impedance of the power supply), the electrons are practically all attracted to the B supply and hardly any to the capacitor. The main reason for this is that the impedance of the B supply is much less than that of the capacitor circuit at the operating frequency; furthermore the positive side of the B supply offers such a ready attraction for the flowing electrons. Consequently something is definitely needed to prevent most of the electrons from flowing to the B supply because only a negligible amount will flow to the capacitor to be amplified further by the following stage.

This undesired feature can be corrected by placing a high impedance in parallel with the capacitor so that the main flow of electrons will pass (in effect) through C_b to tube V_2, and a negligible amount passed to the B supply. This new source of impedance is a resistor and it is placed in the circuit as shown in Fig. 12C. This resistor R_b, having a high value compared to the reactance of C_b, limits the flow of the electrons from tube V_1 to the B supply and therefore, most of them are passed (in effect) through the capacitor C_b to the grid of the tube V_2 to be amplified further. This resistor is known by a number of different names: the "plate resistor, plate-load resistor, or load resistor"; and the capacitor is called either a "coupling" capacitor or a "blocking" capacitor. The reason for this nomenclature of the "coupling" capacitor is evident from the previous discussion of how C_b (in effect) offers a path to, or couples, the electrons from tube V_1 to tube V_2.

If this flow of electrons is alternating in character, capacitor C_b will pass it on to tube V_2. In other words, C_b will only pass current that is alternating and not a direct current. It is, therefore, evident that C_b, in addition to "coupling" the alternating components of the plate current, also "blocks" the flow of any direct current from tube V_1 to tube V_2, and that is why it is sometimes called a "blocking" capacitor. Assuming a resistor to be principally of a purely resistive nature and containing practically negligible inductive and capacitive

characteristics, it can pass both alternating and direct current. In the present case, the resistance R_b has a high value and therefore the a-c component of the plate current will be hindered from flowing through it and pass on to tube V_2 through C_b. This is so because the reactance that C_b offers is much less than the resistance of R_b. All the d-c component of the plate current of tube V_1 has to flow through resistance R_b, regardless of the value, because that is the only path available for it. The value of R_b in itself actually has the effect of limiting the d-c plate current by the simple application of Ohm's Law.

When no connection is between the grid and cathode as in tube V_2 of Fig. 12C, then the grid is said to be disconnected with respect to d-c voltages and currents and is termed a "floating grid". Some means must be provided for the leaking of these extra electrons off the grid to assure plate current flow in tube V_2. This is accomplished by placing a high resistance R_c in series with a bias battery in the grid circuit of V_2 as shown in Fig. 12D. The common name for such a resistor is a "grid-leak" resistor or just a "grid" resistor, and it is placed in the circuit between the grid and the bias voltage supply. If the bias supply in resistance-coupled amplifiers is obtained from self-biased cathode resistors, then the circuit connection will be that shown in Fig. 12A. The usual value of grid resistor is anywhere from 100,000 ohms to about 10 megohms, depending upon the tubes used and other circuit parameters.
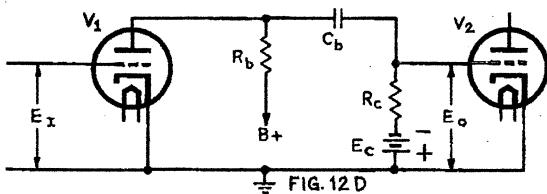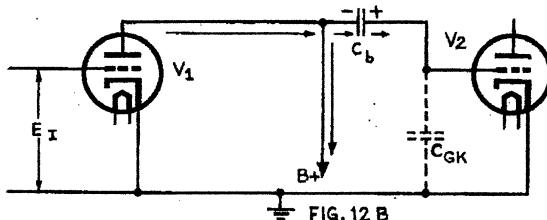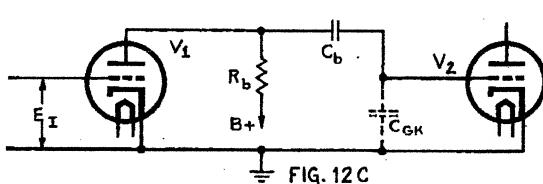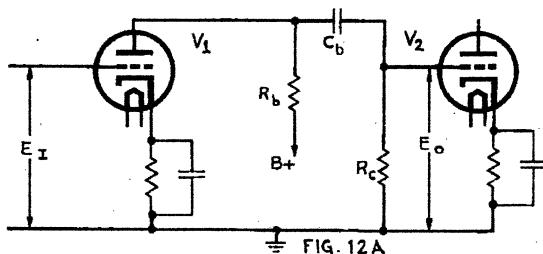

FIG. 12.—These circuits illustrate the development of the R-C coupled network.

Up to this point the need has been considered for the plate load resistance R_b, the coupling capacitor C_b, and the grid resistor R_c, but the reason for choosing special values for them remains to be explained.

(A) The Plate-Load Resistance R_b

Since the value of this parameter and all others are dependent on the tubes used, then it is necessary to discuss them in reference to amplifiers. R_p represents the plate resistance of the tube and its value for triodes is somewhere in tens of thousands of ohms, while for pentodes it is much larger, approaching well into the order of a megohm and higher. What was previously said about R_b was that it had to be large in value, but that was mainly in reference to the signal voltage. In reference to the d-c voltage, the smaller the value the better, but there must be a compromise and that will be discussed now.
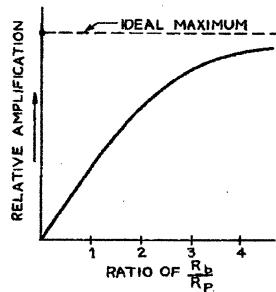

Fig. 13.—This graph illustrates how the ratio of the plate-load resistance to the plate resistance of the tube varies with respect to the relative amplification.

Fig. 13 is a curve showing the way the relative amplification varies as the size of the plate load resistor R_b is increased (shown as a ratio value to the plate resistance R_p of the tube). When the resistance R_b is small, most of the a-c voltage drop is in the plate resistance of the tube itself and the relative amplification is very low. When R_b is made very large, most of the voltage drop will appear across R_b and the relative amplification will increase.

In order to justify the above, it may be thought the plate-load resistance should be about fifty times as large as the plate resistance of the tube. Theoretically this seems sound, but the fact must be considered that R_b is also in the d-c circuit of the tube as well as in the a-c circuit, and the value of R_b must be considered from that angle too. The actual d-c voltage on the plate of the tube is the supply voltage less the d-c voltage drop in R_b. If R_b is quite large, then the d-c voltage

drop in R_b will be large and the actual voltage on the tube will be small — too small to make the tube operative under the existing conditions.

One way to offset this may be to increase the supply voltage to overcome the difference, but such a practice would be wholly impracticable, because the supply would have to produce voltages that approach the order of thousands of volts. Circuits producing such high voltages would be too expensive and also these voltages would not be too conveniently obtained in receivers of today. It is also often impracticable, because the plate load resistor would have to be unduly large in order to dissipate the amount of heat required due to the $I^2_pR_b$ loss. Therefore, a compromise must be obtained between a high value of R_b for a-c amplification and a low value of R_b for d-c operation. The range of such a resistor actually used varies with the type of tube. The plate resistance of triodes is not so large, as mentioned before, and the ratio between R_b and R_p can be anywhere from one to ten. The plate resistance of pentodes is quite high compared to triodes and there are limitations as to how high the load resistance can be, as mentioned above. Consequently it is not practical to obtain too high a ratio between R_b and R_p in the case of pentodes.

(B) The Grid-Leak Resistor R_c

Referring to the diagram of Fig 12D it is understood so far that the total signal voltage drop across R_b will be applied to the grid of the tube V_s through capacitor C_b. This is not really true although it is desired that such a condition exist. Similar to the preceding case in the choosing of a plate load resistor, there are limitations here too. First of all, the capacitor C_b in series with R_c together make up a parallel circuit with R_b to the distribution of signal voltage. Therefore the same signal output voltage that is across the one half of the parallel circuit R_b is impressed across the second half of the parallel circuit, namely C_b in series with R_c. Since the grid of the next tube is connected to the junction of the C_bR_c combination, then this combination acts as a voltage divider of the impressed signal, with the portion of the voltage across R_c being impressed on the grid of the tube. The greater the value of R_c as compared to the reactance of C_b, the

larger will be the voltage impressed on the grid of tube V_s.

Referring to Fig. 14, it is seen that the total signal voltage across R_b, is impressed across the dividing network of R_c and C_b. The voltage that is impressed on the grid will be that across the grid-leak resistor R_c. The greater the resistance of R_c, as compared to the reactance of C_b, the more signal voltage will be impressed on the grid of tube V_s; therefore, why not let the grid-leak resistance be as large as possible, approaching hundreds of megohms? But here the limitation becomes apparent. If this grid-leak resistor is made too large, approaching hundreds of megohms, then the circuit between the grid and cathode can be considered as being practically open circuited. In other words, there is the virtual effect of having the grid "floating", as previously mentioned, because any charge that collects on the grid cannot leak off fast enough due to such high resistance. Stated again, the value of a grid-leak resistor varies anywhere from 100,000 ohms to 10 megohms; depending upon the value of the
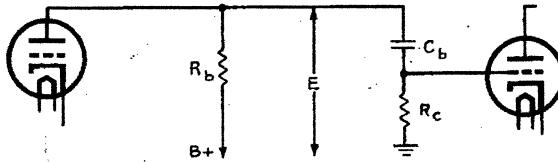

Fig. 14.—An R-C coupled network in which the capacitor-resistor combination, C_b-R_c, acts as a voltage divider.

coupling capacitor and the tubes used. It is quite common, for quick calculations, to choose the grid-leak resistor in proportion to the plate-load resistance (if it is known) and it is usually made about four times as large as R_b. Since most tubes have a value of R_c designated by the manufacturer, then it is sometimes said that the value of R_b should not exceed ¼ of the rated value of R_c, but should exceed the plate resistance of the tube it loads by several times. The aforementioned statement holds true for triode circuits and some pentode circuits, and depends upon the plate resistance of the tube used.

(C) The Coupling Capacitor C_b

Reviewing the function of the coupling capacitor C_b, it is to keep the d.c. from the grid of the second tube and to allow the signal volt-age amplified in the first tube to be coupled to the second tube; or in more concise terms, to offer a low impedance to the signal voltage.

A few considerations in the choice of a coupling capacitor have to be taken into account. It must offer a minimum impedance to the signal voltage, therefore the capacitor will have to be large. That is, seen from the capacitive reactance (the actual impedance offered), $X_{cb} = \dfrac{1}{2\pi f C_b}$, the larger C_b becomes, the smaller X_{cb} becomes. Now if a charge collects on the grid of the second tube, it will have to leak off through R_c, the grid-leak resistor; but if the reactance of C_b is low, this collecting charge will tend to charge C_b, because the resistor R_c offers too high an impedance as compared to X_{cb} and the charge cannot leak off too rapidly. This is undesirable, since it will develop a very high bias on the grid. What we now have is the effect of the impedance of both the capacitor C_b and resistor R_c on this accumulating charge. What must be done under such circumstances is that C_b be discharged under a specified *time limit* before any more charge accumulates on it and therefore to allow passage of the signal voltage to the grid of the second tube. This discharging must naturally occur through R_c, the only available resistor at the input to the second tube.

The Time Constant

The time limit is a mathematical quantity and it is known as the "time constant" of the circuit. It is the product of the coupling capacitance in farads and the grid-leak resistance in ohms (or the coupling capacitance in microfarads and the grid-leak resistance in megohms) and it is expressed in seconds. Mathematically we have:

Time Constant $= R_c C_b$ in seconds.

This product of the resistance and capacitance represents the time required to discharge the capacitor C_b through the resistor R_c to approximately one-third of the original charge (the actual value is 1/2.718).

As an example, assume that it is desired to have a time constant of .004 second with a grid-leak resistor of 2 megohms and the value is wanted of the coupling capacitor to use. Using the above formula,

Time Constant $= R_c C_b$

.004 sec $= 2$ megohms $\times C_b$ microfarads

$$C_b = \frac{.004}{2} = .002 \text{ microfarad}$$

Therefore it is found that the value of the coupling capacitor required is .002 microfarad.

In the use of coupling capacitors certain precautions must be observed. A primary consideration is that the coupling capacitor should not be "leaky" at all. If it has some leakage, then its leakage reactance should be exceedingly high in comparison with the values of plate-load resistance and grid-leak resistance. It should approach the order of thousands of megohms, because, if not, it may be low enough to provide a path for some direct current from the first tube to flow to the grid of the next tube. This would be exceedingly undesirable as it will offset the original bias conditions for operation of the second tube.

GAIN MATHEMATICS

The two most general types of coupling used in radio receivers are transformer coupling and resistance-capacitance coupling. In many of the receivers it is very important to know the stage-gain measurements from any particular stage to the one following, since such information will help the serviceman in analyzing some of the troubles encountered in his work. As has been shown in the past, such measurements can be quickly and accurately made by means of a Chanalyst type of instrument. However, stage gains can also be *calculated* from some known factors about the component parts in the set, as in the case of transformer and R-C coupling. Gain measurements are based upon the input and output voltages of the stages and mathematically the gain is given by the ratio of the output voltage divided by the input voltage. This means that the output voltage is the product of the input voltage and this ratio.
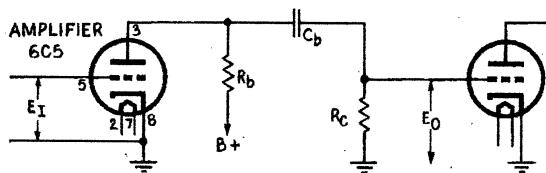
Fig. 15.—A typical R-C coupled amplifier stage in which triodes are employed.

This is usually shown on manufacturers' schematics giving gain-per-stage measurements, as some number followed by X. For example, 20X means that the output voltage of a certain stage is 20 times greater than the input voltage to that stage.

The gain for any stage varies in accordance with the frequency at which it is measured. For instance, if the gain of an r-f R-C coupled stage in the broadcast band is measured, a different value of gain will be found for the low-, middle-, and high-frequency regions of the band. However, for most general purposes the gain, when measured at the middle range of frequencies, may be considered as an overall value for the gain of the stage in question, since the gain at these frequencies is a maximum. Hereafter when a computation is made for the gain of any coupled stage it is to be understood that it is made in the middle range of frequencies.

Gain of an R-C Coupled Amplifier

In Fig. 15 is shown a typical schematic of an R-C coupled triode amplifier using a 6C5 tube. From the diagram and the above definition of gain, it is noted that

$$A = \frac{E_o}{E_i}$$

where A equals the gain of the stage

E_o equals the output voltage measured across R_c

E_i equals the input voltage across the 6C5 control grid

For any R-C coupled triode amplifier similar to that in Fig. 15, the gain is given by the following expression:

$$A = \frac{\mu R_b}{R_p + R_b}$$

where μ equals the amplification factor of the tube receiving the input voltage. (This may be found in any tube manual)

R_b equals the plate load resistor across the amplifier

R_p equals the plate resistance of the amplifier (also found in any tube manual.)

Looking up the characteristics of the 6C5 tube, it is found that under certain operating

conditions (with 250 volts on the plate and —8 volts on the grid), the amplification factor, μ, equals 20 and that the plate resistance R_p under the same operating conditions equals 10,000 ohms. Now if the load resistance R_b across the plate circuit of the tube is equal to 50,000 ohms, the computed gain of the stage would be approximately:

$$A = \frac{20 \times 50,000}{10,000 + 50,000} = \frac{1,000,000}{60,000} = 16.7$$

The gain, A, is therefore 16.7 times as great as the input signal for the particular case under consideration. An important point of interest is that the gain of a triode amplifier can never exceed the rated μ of the tube and can only ideally approach it.
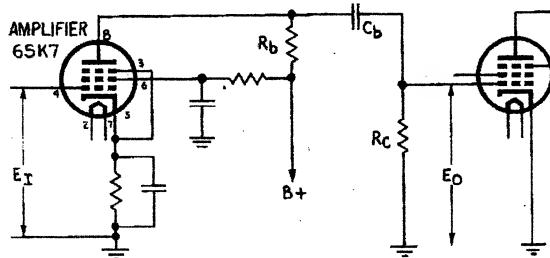
Fig. 16.—A typical R-C coupled amplifier stage using pentodes.

Likewise for a typical R-C coupled pentode amplifier, shown in Fig. 16, the gain is calculated by the following expression:

$$A = G_m R_b$$

where G_m equals the transconductance of the tube in mhos and

R_b equals the plate load resistance of tube in ohms.

The above relation is given under the general assumption that the grid-leak resistor R_c and the plate resistance of the tube R_p are much larger than the plate-load resistor R_b. For the circuit in question, the G_m of the 6SK7 tube is equal to 2000 micromhos (under the operating conditions of 250 volts on the plate, 100 volts on the screen, and —3 volts on the control grid). Therefore with a plate-load resistance on the 6SK7 tube of 100,000 ohms, the gain of the pentode amplifier stage would be approximately:

$$A = .002 \times 100,000 = 200$$

Gain of a Transformer Coupled Amplifier

For transformer coupling there are also simple equations for computing the gain, but the factors involved may be somewhat difficult to find in certain radio sets. One of the primary factors involved in calculating the gain of untuned transformer-coupled stages is the turns ratio of the transformer in question. Before going on, it would be best to try to understand ways of finding out the number of such turns. First, such information may be given in the service data sheets; second, it may be an easy matter to count the number of turns on certain coils; and third, the only alternative is to try to obtain such information directly from the manufacturer.

The simplest type of transformer-coupled gain calculations are made with a transformer-coupled stage where the primary and secondary are untuned. Under such conditions the gain, A, is given by the following relation,

$$A = \frac{E_o}{E_I} = \mu \, n$$

where μ equals the amplification factor of the tube in question, and

n equals the turns ratio of the secondary to primary. The gain can be easily calculated, when the turns ratio is known.

On the other hand, however, a great number of transformer-coupled stages are found in radio receivers that have either one or both of the transformer windings tuned. Under these circumstances the calculation of gain becomes somewhat more difficult and it involves too much mathematics to be considered here. It involves the usage of the different impedance values of the primary and secondary along with the reflected impedance.

It is important at this time to understand a few facts about the stage-gain measurements as supplied by the manufacturers with their service data. Generally speaking, these gain measurements are *not* supposed to be considered as standard values for every set of the same model put out by the manufacturer. In other words, if there are three radio sets, all the same model supplied by a certain manufacturer, the gain measurements in only one of the receivers may be near those listed in the service data supplied for this model. This does not mean that the lower stage gains of the other sets indicate that they are not working properly. Most manufacturers select an *average* of the gain

measurements made on quite a few sets of the same model and publish these figures as applicable to a certain model. The gain readings then represent not only one set but the average of values among them all. Under these circumstances the gain readings that are made on a set may be above or below those given by the manufacturer and still be satisfactory. Only if the stage gain differs appreciably from that given on the service data should the servicemen then question the functioning of the set.

Motorola Model 39B-2

In the Galvin Motorola Model 39B-2 an untuned r-f transformer is used to couple the r-f signal from the output of 6SK7GT r-f amplifier to the signal grid input of the 6SA7GT oscillator-modulator tube, as shown in Fig. 17. The transfer of energy from one circuit to the other is carried out by the means of ordinary transformer action. Due to the close coupling between the primary and secondary of the transformer, a certain amount of mutual inductance is offered and consequently the desired amount of energy is fed from one circuit to the other. A very unusual fact about this circuit is that the r-f transformer used is completely untuned. In other words, the primary and the secondary circuits are not tuned and such a transformer should have a very broad selectivity.

The input to the 6SK7GT r-f amplifier and the oscillator are tuned by ganged capacitors. Therefore in order to *follow* the tuning of these capacitors throughout the 545-1600-kc frequency range of the broadcast band, the r-f transformer, part (2), has to pass the entire band of broadcast frequencies. The frequency-response characteristics of this type of transformer is such that there is a decrease in response at both the high- and low-frequency

ends of the broadcast band; therefore, when tuning in different stations the selectivity of the set probably fell off at either end of the tuning range.

In order to make the selectivity of the set broader and not have this falling-off in response at both ends of the band, a 33,000-ohm resistor was placed across the primary and secondary of the r-f transformer in question, as shown in Fig. 17. These resistors reduce the overall Q of the transformer and consequently lower the gain of the stage. By reduction in the Q and lowering of the gain, the response curve is made effectively broader and the transformer will then pass all the frequencies of the broadcast band. Even though the overall gain of the stage is lowered, there will still be enough gain in this stage for the proper functioning of the receiver.

One of the benefits secured by this type of untuned r-f coupling is that only two ganged capacitors are used, thereby constituting an economical saving by doing away with the necessity of another ganged capacitor section for tuning. Since space limitations in radio sets are always of importance, this type of coupling saves a certain amount of vitally needed space.

Admiral Model 6E1

In the Admiral portable models 6E1 and 6E1N a single tuned stage is used to couple the signal from the 1N5 r-f amplifier to the signal grid of the 1A7 converter tube, the diagram of this circuit being shown in Fig. 18. The single-tuned circuit consists of a 420-mmf fixed mica capacitor $C3$ in parallel with a permeability tunable coil $L3$. In this type of circuit there is no transformer action since only one coil is used. The purpose of this type of coupling is to offer a maximum load impedance to the 1N5 r-f amplifier and, therefore, get the maximum gain.

It will be recalled, that the impedance of a parallel tuned circuit is a maximum at resonance, and it is purely resistive in nature. Also, according to the gain equations of pentode amplifiers, the gain of such a stage is approximately equal to the G_m of the amplifier tube multiplied by the effective load on the tube. Since the load on this tube at resonance is a maximum and purely resistive in nature, as stated above, then the gain of the stage is dependent upon this load. The impedance of a parallel tuned circuit of a fairly high Q at res-
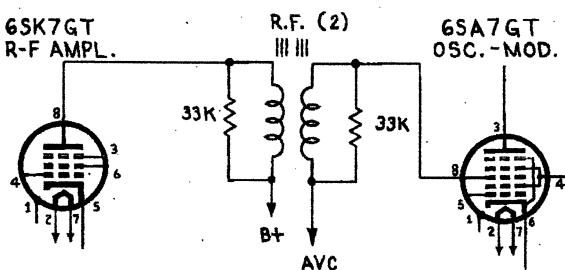

Fig. 17.—Untuned transformer coupling is used in the r-f stage in the Motorola Model 39B-2.

onance is equal to $(X_L)^2$, where X_L is the in-

$$\frac{}{R}$$

ductive reactance of the coil in the circuit and equal to $2\pi f L$ and the resistance R is equal to the resistance of the coil in the circuit. Since the Q of the coil is equal to the ratio of the inductive reactance to the equivalent series resistance of the coil, then the impedance of a parallel resonant circuit can also be expressed as $X_L Q$. The gain of the stage in question is then equal to the following:

$$A = G_m \ \frac{(X_L)^2}{R}$$

or

$$A = G_m X_L Q.$$

In either equation G_m is equal to the transconductance of the 1N5 tube expressed in mhos and X_L, R, and Q are defined as above. Since the capacitor of the tuned circuit is a mica capacitor and as such has negligible losses, then the Q referred to can be considered as the Q of the coil. The gain of this stage can be easily calculated from either one of the above equations provided all the factors are known.

For this circuit of Fig. 18 the G_m of the 1N5 tube is equal to 750 micromhos, the resistance R of the circuit can be measured between the high and low side of the tuned circuit, and the inductive reactance X_L can be calculated from the frequency of operation. The method of obtaining the inductive reactance without knowing the actual inductance of $L3$ is as follows. The resonant frequency of the tuned $L3$-$C3$ circuit is read directly from the station selector dial. The inductive reactance of $L3$ at resonance is equal to the capacitive reactance of $C3$. Since we know the value of $C3$ (420 mmf), the capacitive reactance of the circuit can be calculated and this value substituted for X_L in the above equations. Tune the station selector dial to a frequency of 710 kc and assume that the d-c resistance measured across the tuned circuit (with the set not in operation) is equal to 10 ohms, then the gain of the stage may be computed as follows:

At resonance the inductive reactance of the coil $L3$ is equal to the capacitive reactance of the capacitor $C3$ and therefore

$$2\pi f(L3) = \frac{1}{2\pi f(C3)}$$

with the capacity expressed in farads and the

frequency in cycles per second. In the circuit under consideration f equals 710 kc and $C3$ equals 420 mmf. Consequently the capacitive reactance is equal to

$$X_C = \frac{1}{2 \times 3.14 \times 710 \times 10^3 \times 420 \times 10^{-12}}$$

$$= 534 \ ohms$$

therefore, the inductive reactance X_L is equal to 534 ohms, since $X_L = X_c$.

The gain of the stage will be equal to

$$A = G_m \ \frac{(X_L)^2}{R}$$

$$= .000750 \times \frac{(534)^2}{10} = 21.4$$

The gain of this stage at 710 kc and with the d-c resistance of the tuned circuit assumed to be equal to 10 ohms and also to be equal to the r-f resistance, and with G_m expressed in mhos, the computed gain is equal to 21.4.
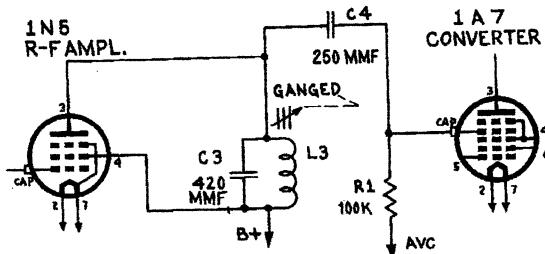
Fig. 18.—The single-tuned r-f coupling in the Admiral Model 6E1 and 6E1N allows for maximum gain and good selectivity in this stage.

Since the gain analysis is based on the mathematics of R-C coupled networks, the question might be raised of substituting as large a resistor as possible in the plate circuit of the 1N5 r-f tube as a load. The primary purpose of inserting the parallel tuned circuit into this coupling arrangement, is to obtain a high degree of selectivity, which is more important than achieving a maximum amount of gain in this circuit.

Zenith Model 6G001

In the Zenith Model 6G001 an interesting transformer coupling arrangement is used between the output of the 1LN5 r-f tube and the signal-grid circuit of the 1LA6 converter tube. The r-f transformer used, shown in Fig. 19, has its primary untuned and its secondary

tuned, which is common to such types of coupling. However, the interesting fact about this coupling arrangement is that the signal-grid circuit of the converter tube is *tapped* to the secondary inductance instead of going directly to the high side of this coil. The circuit components used in this receiver necessitated the tapping down of the coil in order to obtain proper high-frequency tracking, proper band coverage, and a high L-to-C ratio of the tuned circuit.

For the moment let us consider just the tuned secondary circuit of the r-f transformer $L2$ in Fig. 19. This tuned circuit consists of the transformer secondary coil L_s in parallel with the ganged tuning capacitor $C1$ and the trimmer $C7$. The tuning capacitor $C1$ has the inherent quality of containing too high a minimum amount of capacitance. With the capacitor plates fully out of mesh a certain amount of capacitance exists (which is an inherent quality of *any* tuning capacitor) and this capacitance is referred to as the minimum capacitance that a tuning capacitor possesses.

It so happens that the minimum amount of capacitance that $C1$ offers is too high and due to this, improper tracking at the high frequencies results. This high minimum capacitance also produces a lower L-to-C ratio for the tuned circuit. This low ratio is undesired as it causes improper band coverage over the tuning range of frequencies in this set. By causing improper band coverage is meant that the tuning range of capacitor $C1$ is not great enough to cover the broadcast frequency range of this set which is 535-1620 kc. All of these

Fig. 19.—The signal grid of the 1LA6 converter is connected to a tap on the tuned secondary of the r-f transformer in the Zenith Model 6G001.

defects are compensated for by increasing the L-to-C ratio of the tuned circuit. Without changing component parts it may at first appear difficult to raise the L-to-C ratio. However, if it is realized that every tube has intrinsic qualities, such as input and output capacitances beside the interelectrode capacitances between tube elements, it will be understood how the L-to-C ratio is increased.

The input capacitance of the 1LA6 converter tube is effectively between the signal grid and ground and therefore this input capacitance would exist across the complete tuned circuit if the signal grid circuit were attached to the high side of coil L_s. If a tube manual is examined, it will be found that the input capacitance of the 1LA6 converter tube is equal to about 7.5 mmf. This capacitance may not affect the tracking at the low-frequency end of the band, but this input capacitance will have appreciable effect upon the tracking at the high-frequency end of the band usually by causing improper tracking at these frequencies.

The method of *effectively reducing* this input capacitance across the tuned circuit is accomplished in this set by taking the 1LA6 converter signal-grid voltage off a *tap* on the tuned circuit inductance L_s. By tapping down on this coil, the L-to-C ratio is increased and by this increase the desired band coverage and proper high-frequency tracking is attained.

The way proper high-frequency tracking is attained is as follows: Originally with the trimmer $C7$ across $C1$, there was a certain amount of overall minimum capacitance at the high-frequency end of the band. This minimum capacitance included the input capacitance of the tube, which is quite high as noted. It was consequently found that when trying to use the trimmer $C7$ to track at the 1400-kc high-frequency end of the band, there was too much capacitance in the circuit to get proper tracking. With the effect of the input capacitance greatly reduced, the total effective minimum capacitance across the tuned circuit is likewise reduced and proper tracking at the 1400-kc high-frequency end of the band is attained.

The high L-to-C ratio is realized at once, because the overall total effective capacitance in the parallel tuned circuit is reduced. This, as mentioned before, brings about greater band coverage. How this coverage is increased can best be explained by the following analysis.
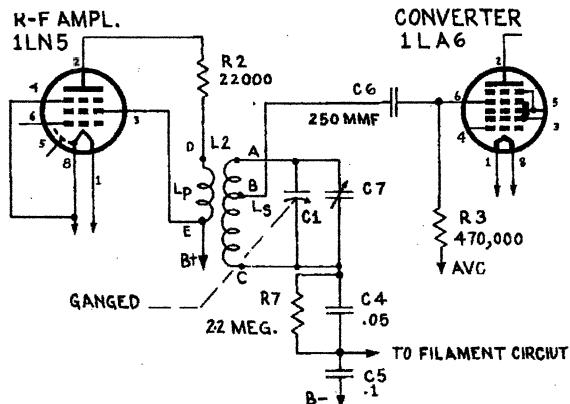
It has been mentioned that by greater band coverage is meant covering the entire frequency range of the set with the variable ganged tuning capacitor. In other words, with a *high* L-to-C ratio, when the tuning capacitor is varied by a certain amount, a greater frequency spread is covered on the tuning dial than if a lower L-to-C ratio was used. This can be illustrated by examination of the resonant-frequency formula $f = \dfrac{1}{2\pi\sqrt{LC}}$ and applying arbitrary values of L and C to prove the above statements. Let us assign arbitrary values of 100 microhenrys for L and 20 mmf for C and consider this as a high L-to-C ratio. If C is varied 10 mmf, that is, if C is varied first to 15 mmf and then to 25 mmf, the difference in frequency change would be as follows:

for $C = 25$ mmf, $f = \dfrac{1}{2\pi\sqrt{LC}}$

$$= \frac{1}{2 \times 3.14 \times \sqrt{100 \times 10^{-6} \times 25 \times 10^{-12}}}$$

$= 3.18$ mc.

for $C = 15$ mmf, $f = \dfrac{1}{2\pi\sqrt{LC}}$

$$= \frac{1}{2 \times 3.14 \times \sqrt{100 \times 10^{-6} \times 15 \times 10^{-12}}}$$

$= 4.11$ mc.

The frequency range covered by this tuning is 4.11 mc. less 3.18 mc. which is equal to 930 kc. Now keeping the original LC *product constant* and making the L-to-C *ratio lower* by making L equal to 50 microhenries and C equal to 40 mmf, let us examine the change in frequency when the capacitor C is again varied 10 mmf; that is, C is varied between 35 mmf and 45 mmf. Therefore at the extremes of 35 mmf and 45 mmf we find that:

for $C = 45$ mmf, $f = \dfrac{1}{2\pi\sqrt{LC}}$

$$= \frac{1}{2 \times 3.14 \times \sqrt{50 \times 10^{-6} \times 45 \times 10^{-12}}}$$

$= 3.36$ mc.

for $C = 35$ mmf, $f = \dfrac{1}{2\pi\sqrt{LC}}$

$$= \frac{1}{2 \times 3.14 \times \sqrt{50 \times 10^{-6} \times 35 \times 10^{-12}}}$$

$= 3.80$ mc.

The frequency range covered by this type of tuning that has a lower L-to-C ratio, is 3.80 mc less 3.36 mc, which is equal to 440 kc.

In comparing the above calculations it will be seen that a higher L-to-C ratio causes a greater change in frequency for a given change in the tuning capacitor than does a lower L-to-C ratio. This was shown under the provision that the product of LC is constant to start with.

If the diagram of Fig. 19 is examined once more it will be noted that coil L_s is tapped down only about one-third of the way from the high end. It would appear that further lowering the tap on L_s would give still better band coverage, which is true. However, there is a limitation to how far down the coil can be tapped and this limitation is due to the fact that the lower down the coil L_s is tapped, the lower will be the applied r-f voltage to the signal grid of the 1LA6 tube. In other words, all the signal voltage appearing across coil L_s is not being used and the r-f voltage fed to the signal grid of the 1LA6 tube, is controlled by the tap on L_s.

Tracing the path that the r-f signal energy takes in going from the plate of the 1LN5 tube to the signal grid of the 1LA6 converter is as follows: The r-f signal appearing across the primary L_p of the r-f transformer $L2$ induces a voltage across the complete secondary coil L_s. Part of this r-f induced voltage is tapped off L_s from points B (which is about one-third down from the high side of L_s) to C and is applied onto the signal grid of the 1LA6 converter tube through the 250-mmf coupling capacitor $C6$.

Globe Model 62

In most radio sets where two i-f amplifier stages are employed, three i-f transformers are used. One is used as the input i-f transformer, the second as the intermediate i-f transformer, and the third as the output i-f transformer. However, in the Globe Model 62 two i-f amplifier stages are used, but only *two* i-f transformers are in the circuit. R-C coupling is used between the two amplifiers instead of the usual

intermediate i-f transformer, the circuit for this arrangement being shown in Fig. 20. In order to get appreciable r-f gain, most sets have an r-f tube in the input section. In this receiver an extra r-f stage is not used to obtain the increase in gain, but an extra i-f amplifier is used. By the addition of this extra i-f amplifier stage, the added gain required is obtained.

One advantage of using an extra i-f amplifier instead of a separate r-f amplifier is that the extra tuning of the r-f stage is eliminated and at the same time the extra gain that is needed is available. The reason why an intermediate i-f transformer is not used is due to the fact that all the *necessary* gain and selectivity is accomplished by using R-C coupling between the i-f stages. In other words, this receiver needed extra gain in order to have it operate satisfactorily and if an intermediate i-f transformer was used more than enough gain would be available. In Fig. 20, it is shown
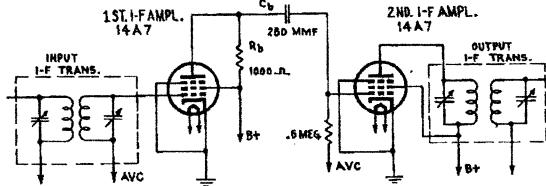

Fig. 20.—In the Globe Model 62, R-C coupling is used between the two i-f amplifiers instead of the usual i-f transformer coupling.

that the load resistance R_b in the first i-f tube plate circuit is kept low (1000 ohms), so as not to have too great an increase in gain. From the gain mathematics of an R-C coupled pentode amplifier we know that the gain A is:

$$A = G_m \times R_b$$

From tube manuals, G_m for the 14A7 tube, under operating conditions of about 100 volts on the plate and screen and with zero bias voltage on the control grid, is about 2100 micromhos. Therefore the gain will be

$$A = .0021 \times 1000 = 2.1$$

From this value of gain it is seen that only a small amount of amplification is needed and this is secured by R-C coupling of the two i-f stages.

Zenith Model 7ML080

In the Zenith Model 7ML080 a combined coupling arrangement is used between the 7A7 r-f amplifier and 7B8 converter. This coupling arrangement consists of the regular

R-C coupling combined with single tuned-circuit coupling. The schematic for this coupling arrangement is shown in Fig. 21. The primary
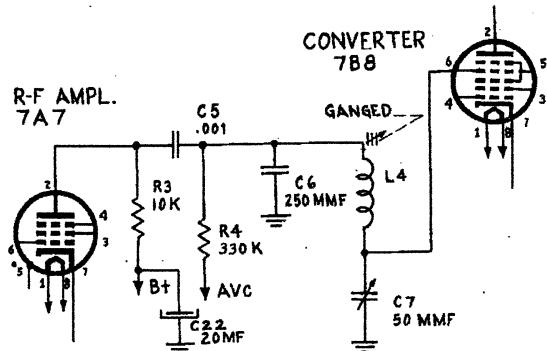
Fig. 21.—The coupling between the r-f and converter tubes in the Zenith Model 7ML080 consists of the regular R-C coupling combined with single-tuned circuit coupling to obtain proper selectivity and better image rejection.

purpose of utilizing this coupling system is to obtain good selectivity of the r-f signal and better image rejection.

In comparing the overall frequency response characteristics between transformer coupling and R-C coupling, it is a known fact that transformer coupling, as well as tuned-circuit coupling, allows for greater selectivity but not a great band width. R-C coupling on the other hand has a very broad response compared to transformer coupling and it is therefore not as selective. With the differences between these two types of coupling established, it appears as though the effects of both are simultaneously present in the circuit of Fig 21. The R-C coupling arrangement, consisting of $R3$, $R4$, and $C5$, allows for broader band coverage, and the tuned-circuit coupling, consisting of $L4$, $C6$, and $C7$, produces good selectivity. Combining both of these arrangements, however, introduces certain properties into this circuit that are not readily evident.

The combination of $L4$, $C6$, and $C7$ is in effect a parallel tuned circuit. The two capacitors are effectively in series (through ground) and they both are considered to be in parallel with $L4$. The interesting part of this circuit is that by taking the output signal voltage across $C7$, the resonant circuit is made to function as a tuned autotransformer. Usually, in tuned-circuit coupling the signal grid of the converter tube is attached to the high side of the tuned circuit, as is the output of the r-f amplifier or the antenna. This contributes to the simplicity

of the coupling, and thereby reduces the possible sources of trouble. On the other hand, this simplicity has its drawbacks as well as its advantages, and the reason for the connection used here will be explained below.

If the ground point was removed from the junction of C6 and C7 and placed at the junction of L4 and C7, then R3 would be effectively in parallel with the entire tuned circuit. This is so since R3 is connected to the high end of the tuned circuit through the .001-mf capacitor C5, which has a very low reactance in the broadcast band; for example, it is only 159 ohms at 1000 kc. The low end of R3 is connected to ground, as far as r-f is concerned, through the power-supply output-filter capacitor (C22) connected from it to ground. Because of the loading effect that R3 would have, the Q of the tuned circuit would be decreased. This would increase the band width; putting it differently, this would decrease the selectivity.
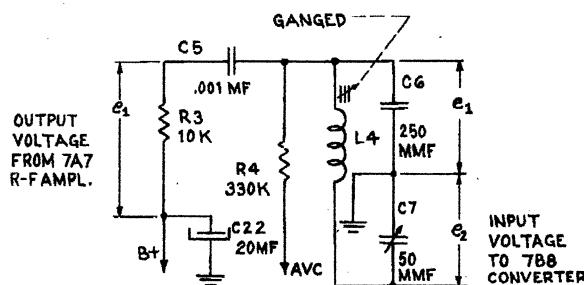

FIG. 22.—Simplified schematic of the coupling arrangement of Fig. 21.

Returning to the actual circuit, a better idea of its operation may be had by redrawing it as in Fig. 22. Designate e_i the output voltage of the r-f amplifier. Then the voltage across C6 will be virtually equal to e_i. As was pointed out in the preceding paragraph, the high ends of R3 and C6 are connected through the very low reactance of C5. In addition, the low ends (ground end of C6) are connected through the negligible reactance of the output capacitor (C22) of the power supply. Now consider the tuned circuit. It consists of an inductive reactance L4 and a capacitive reactance (C6 and C7 in series) in parallel, with a tap on the capacitive arm. In the usual tuned autotransformer the tap is found on the inductive arm; this is the only difference between the usual circuit and the one in Fig. 22, and it is of little importance electrically. Therefore the present

circuit may be considered as a true tuned autotransformer.

(As an analogy to the present case, consider the tank circuits of Hartley and Colpitts oscillators. Both tanks are true autotransformers, which couple the signals necessary to maintain oscillation from the plate to the grid in the proper amplitude and phase relation. In the Hartley tank, the tap is in the inductive arm, while in the Colpitts tank it is in the capacitive arm, as it is in the tuned coupling autotransformer that is shown in Fig. 22. See section on "Oscillators.")

In many respects the operation of an autotransformer in a particular application can be treated like that of a transformer with no direct electrical coupling. For example, we can speak of a primary and a secondary. In the case of the circuit of Fig. 22, the primary terminals are the high end of C6 (connected to C5) and ground; the secondary terminals are the high end of C7 (connected to the signal grid of the 7B8) and ground. It can be proven mathematically that in this circuit the primary and total circuit impedances are related by the following equation:

$$\frac{Primary\ impedance}{Total\ impedance} = \left(\frac{C7}{C6 + C7}\right)^2$$

(It may appear peculiar that C7 instead of C6 is found in the numerator of the squared fraction. The reason for this is that capacitive reactances—which are very important here—are inversely related, that is, the *larger* the capacitance, the *smaller* the capacitive reactance.) Rewriting the equation we obtain:

Primary impedance

$$= Total\ impedance \times \left(\frac{C7}{C6 + C7}\right)^2$$

$$= Total\ impedance \times \left(\frac{50}{250 + 50}\right)^2$$

$$= Total\ impedance \times \frac{1}{36}$$

In other words, by connecting the input across C6 alone, the effective impedance to the input is reduced to *1/36* of the total impedance of the parallel circuit. In this way the loading effect of R3 is very much reduced, for instead of

being parallel with the high impedance of the entire tuned circuit, it is across only *1/36* of this impedance.

A further advantage of this tuned autotransformer is that it provides a voltage step-up. The primary (input) and secondary (output) voltages are *inversely* proportional to the values of the primary and secondary capacitors: that is,

$$\frac{Secondary\ voltage}{Primary\ voltage} = \frac{C6}{C7}$$

$$Secondary\ voltage = \frac{C6}{C7} \times Primary\ voltage$$

$$= \frac{250}{50} \times Primary\ voltage$$

$$= 5 \times Primary\ voltage$$

By obtaining good selectivity, this type of circuit also allows for better image rejection. The intermediate frequency of this automobile receiver is 265 kc (which is lower than the usual 455 kc used in home receivers) and when the i.f. is as low as this, then there is a greater chance for the image frequency to make its way through the selective r-f tuned circuit in sufficient strength to be troublesome. This can be easily illustrated by examining the frequency-response curve (or selectivity characteristic) of Fig. 23.
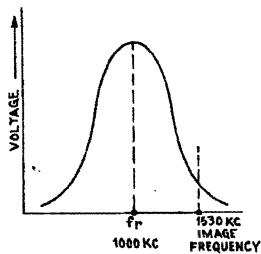
Fig. 23. — Selectivity response curve illustrating image-frequency rejection.

If the set is tuned to an r.f. of 1000 kc, then the oscillator would be tuned to 1265 kc to give an i.f. of 265 kc. If a fairly strong 1530-kc image signal also appeared on the input to the set and made its way through the r-f amplifier stage, it would also beat with the oscillator frequency of 1265 kc to give an i.f. of 265 kc. Therefore the selectivity of the tuned circuit has to be sharp enough at the center frequency of 1000 kc to reject the interfering signal. In

other words f_r in Fig. 23 is the resonant frequency of the tuned circuit and equal to 1000 kc. The band width should be sufficiently narrow that a signal at 1530 kc will not be passed through the r-f stage to any appreciable extent. If the band width is sufficiently narrow, then the selectivity is sharp and interfering image signals are rejected. The greater the intermediate frequency, the less restrictions have to be placed on the selectivity, but as the i.f. decreases the selectivity has to increase in order to have good image-frequency rejection.

HIGH-FREQUENCY COMPENSATION ADAPTED TO BROADCAST-SW RECEPTION IN G. E. MODEL 321 (Early) and ESPEY MODEL RR-13L

In the video-frequency sections of many radar and television receivers the band-width requirements are quite high. In order to pass the proper range of frequencies, the R-C coupling arrangements used, needed a very wide and flat response characteristic. That is, in the ordinary R-C coupling arrangements the flattest response was at the middle frequencies and at the high- and low-frequency end of the band the response curve dropped, i.e. the voltage gain decreased. Consequently, special coupling arrangements, *called high- and low-frequency compensating networks*, were inserted between certain video-frequency stages in order to *peak* up the high- and low-frequency ends of the response curve. The band-width of the video or picture frequencies in television receivers varied from around 20 cycles to as high as 4 megacycles. Such a wide band-width can be easily understood to have a decrease in the high- and low-frequency ends of the response curve with ordinary R-C coupling and that certain compensations are needed to elevate these ends. In compensating for high frequencies, inductances were connected into the circuit in special places so as to peak the high end of the response curve.

In many radio receivers utilizing broadcast and short-wave bands, it has been found that when R-C coupling is used between r-f stages there is a decrease in the high-frequency end of the response curve. Consequently, many of these models contain inductances in conjunction with the R-C coupling already used in

their r-f stages to compensate for the loss in gain (i.e. the drop in the response curve) at the high frequencies in a similar manner to the way it is done in television sets. Typical circuits in which these compensating inductances are employed can be found in the Espey Model RR-13L, Fig. 24, and in the General Electric Model 321, (Early), illustrated in Fig. 25.

In the Espey model, shown in Fig. 24, only a *"series peaking"* coil is used in series with
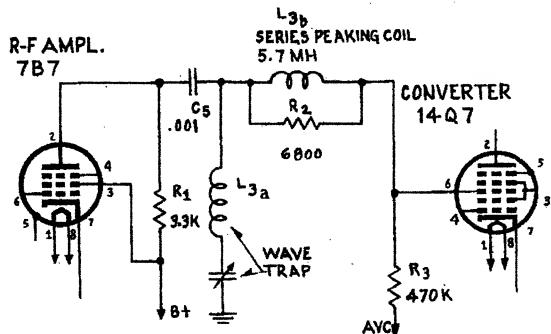
Fig. 24.—A series peaking coil for high-frequency compensation is used in the r-f stage of the Espey Model RR-13L.

the plate-to-grid circuit between the 7B7 and 14Q7 tubes; whereas in the G.E. Model 321 of Fig. 25, besides a "series peaking" coil, a "shunt peaking" coil is used in the plate load circuit of the 12SK7 tube. Before entering into a discussion of these two models, it will be best to analyze high-frequency compensating networks from a generalized viewpoint.

R-C coupling is often used to couple two r-f stages in order to obtain as wide a band-pass characteristic as possible. In many instances
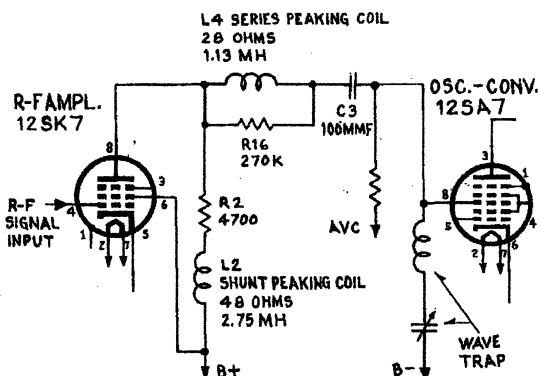
Fig. 25.—Both series and shunt peaking coils for high-frequency compensation are used in the General Electric Model 321 (Early).

it has been found that the high-frequency response of such R-C coupled networks drops and consequently the desired band-width characteristic is unobtainable. The reason a loss in gain occurs at the high frequencies in R-C coupled networks is primarily due to total shunt capacitances existing in the coupled circuit, these being shown in Fig. 26. C_1 is the *effective* shunt capacitance existing at the input to the R-C coupling arrangement and primarily consists of the output capacitance of tube V_1 and the stray wiring capacitances in the vicinity of tube V_1. C_2 is the *effective* shunt capacitance existing at the output of the R-C coupling arrangement and primarily consists of the input capacitance of tube V_2 and the stray wiring capacitances in the vicinity of tube V_2. The stray wiring capacitances include those existing between capacitors C_b and C_s to ground and any other conceivable type of wiring in the coupling circuit that may give appreciable capacitance to ground. These total shunt capacitances, C_1 and C_2, are inherent qualities of any coupling network and they are difficult to reduce. In fact, the only effective capacitance that may be reduced is the stray wiring ca-
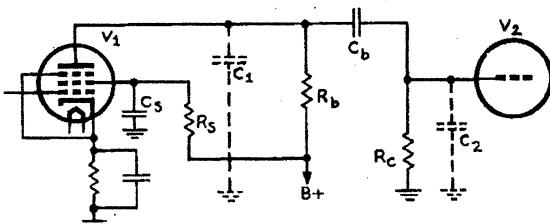
Fig. 26.—A typical R-C coupled stage in which the shunting capacitances C_1 between the plate and ground and C_2 between the grid of the next tube and ground are indicated.

pacitances, but even though the wiring is moved a bit, always some stray wiring capacitance exists which causes troubles at the high frequencies. Added up, these shunt capacitances are the effective elements that reduce the gain of R-C coupled networks at high frequencies. The reactance of these capacitances are such that they offer a low-impedance path to ground at the high frequencies and thus the gain of the R-C coupled stage is reduced.

From the gain mathematics of R-C coupled pentode amplifiers it was shown that the gain of the stage was dependent upon the G_m of the amplifier tube and the effective load resistance, or mathematically:

$$A = G_m R_b$$

For purposes of quick computations it was stated that this equation was the gain at the middle frequency region of the band in question and that it would be considered as being the overall gain of the stage. At this point, however, it should be pointed out that true gain at the high frequencies is less because of the shunting capacitances. The true gain at the high frequencies of any R-C coupled network is given by the following relation:

$$\frac{A_{high}}{A_{mid}} = \frac{1}{\sqrt{1 + \left(\dfrac{R_b}{X_{CT}}\right)^2}}$$

$$= \frac{1}{\sqrt{1 + (2\pi f C_T R_b)^2}} \quad (1)$$

where A_{high} equals the gain at the high frequencies

A_{mid} equals the gain at the middle frequencies and equals $G_m R_b$

R_b equals the resistive load on the plate of the amplifier in ohms

X_{CT} equals $1/2\pi f C_T$, the capacitive reactance of C_T in ohms

C_T equals the total shunt capacities in farads $(C_1 + C_2)$ and

f and π are the same as used in previous explanations.

This equation is given under the condition that the plate resistance R_p of the amplifier and the grid-leak resistor, R_c, of the following stage, are both much larger in value than R_b and can be neglected as being part of the load on the amplifier.

In examining Fig. 26 the question might well be raised, how can C_1 and C_2 be added, as they are not exactly in parallel but separated by capacitor C_b? That is really an interesting question and the answer is the basic reason why inductances are used in special places to compensate for the high-frequency loss in gain.

The answer to this question is based upon the reactance that C_b offers at high frequencies. The average coupling capacitor C_b (as employed in the radio receivers of today) is about 250 mmf when used in an r-f stage, and its reactance at high frequencies is quite low and can be considered as a short between capacitances C_1 and C_2. In fact, at the 1500-kc region

of the broadcast band the reactance of C_b is equal to only 425 ohms. Due to this low value of reactance, capacitances C_1 and C_2 can be considered as being effectively in parallel and thus are added. C_1 and C_2 together are easily seen to offer a ready path to ground for the high frequencies and thus reduce the gain of the stage at these frequencies.

From the above mathematical relation (equation 1) we find that the gain of an R-C coupled stage at high frequencies is some fraction of the gain at the middle frequencies. Rearranging the above expression and substituting $G_m R_b$ for A_{mid}, this becomes

$$A_{high} = \frac{G_m R_b}{\sqrt{1 + (2\pi f C_T R_b)^2}} \quad (2)$$

This relation is in reality not at all difficult to understand. Let us consider equation 2 for a moment. What this relation states (and every mathematical expression is a form of statement) is that the gain of an R-C coupled amplifier stage at high frequencies, A_{high}, is dependent upon four distinct factors, these being the transconductance or G_m of the tube used; the plate load, or R_b, on the same tube; the frequency of operation f, and the total shunting capacitances C_T. The other quantities used in the expression are no more than actual numbers or numerical constants. The expression further states that the true gain of the R-C coupled stage involves the four above mentioned quantities in such a way that the gain is equal to the product of the transconductance of, and the plate load on, the amplifier tube, all divided by the relation inside the square-root sign. This square-root relation is equal to one (1) plus the square of the product of a numerical constant 2π which equals 2×3.14 or 6.28, and multiplied by the frequency of operation f, the shunting capacitances C_T, and the plate load R_b.

Consequently from the above expression it is found that if the G_m of the tube in question is high, so will the gain, A_{high}, be high, but if the shunt capacitances, C_T, of the tube and the frequency of operation, f, are high, then the gain A_{high}, will decrease. It can be seen how much information this mathematical expression gives. Even an actual computation is not difficult, when all the values of the G_m, R_b, f, and C_T are known; it can be easily and simply executed in a few steps. Let us assign values to the sym-

bols and illustrate a typical calculation, which consists of a series of multiplications and a division. Assuming that

R_b equals 50,000 ohms, G_m equals 1000 micromhos, f equals 1500 kc, and C_T equals 30 mmf $(C_1 + C_2)$, then

$$A_{high} = \frac{.001 \times 50,000}{\sqrt{1 + (2 \times 3.14 \times 1500 \times 10^3 \times 30 \times 10^{-12} \times 50,000)^2}}$$

Starting with the denominator (the expression under the square-root sign), we find that $(2 \times 3.14 \times 1500 \times 10^3 \times 30 \times 10^{-12} \times 50,000)^2$ is nothing more than a series of multiplications and is equal to 14.1. Squaring this product simply means multiplying 14.1 by itself, which gives 199. Adding this number to one (1), gives the total number under the square-root sign which is $1 + 199$ or 200. Evaluating the square root of 200 ($\sqrt{200}$), gives that number which multiplied by itself equals 200 and it is found that 14.14 is close enough for practical purposes. Up to this point only the denominator has been calculated, which is equal to 14.14. The numerator of the expression is then equal to $.001 \times 50,000$ or 50. From here on only one more step is required to obtain the final value of A_{high}, and that is to divide the numerator by the denominator. Therefore

$$A_{high} = \frac{50}{14.14} = 3.53$$

From the above mathematical computation the gain at the middle frequencies which is equal to $G_m R_b$, is 50, as compared to the gain at the high-frequency end of the band, which is 3.53. This drop in gain, as mentioned before, is due to C_T. The undesired effect of C_T is reduced by inserting inductances in the circuit at special places. The primary purpose of inserting these inductances is to increase the load impedance of the amplifier effectively at the high frequencies, thereby increasing the gain lost by the effect of the shunting capacitances.

Shunt Peaking

One of the methods of using inductances to increase A_{high} is to insert in series with the plate resistor R_b an inductance which is commonly called a "peaking coil," because it

"peaks" the high-frequency end of the response curve. It is also often called a "shunt" peaking coil, because it is part of the parallel load on the amplifier tube. In Fig. 27 a typical R-C coupled stage is shown which uses a shunt peaking coil, L_p. The shunt capacitance C_1 and C_2 are shown lumped as one capacity, namely C_T for reasons which were previously explained. The effective load impedance on tube V_1 is the combination of R_b in series with L_p, both being in parallel with C_T.

An analysis of this load circuit in the boxed portion of Fig. 27 is necessary in order to show how the effective load impedance is increased in order to increase the gain. R_c is neglected as part of the load impedance because it is usually made very high and therefore it has very little effect as part of the parallel load. This load impedance is nothing more than a parallel resonant circuit with resistance in the inductive arm. From previous explanations of parallel resonant circuits, it was shown that the impedance of such a circuit is a maximum and purely resistive at resonance; consequently, it is seen how the effective load on the tube is increased due to the insertion of L_p. The important point to consider here is how to choose the value of L_p and what frequency should be used to figure the coil to "peak" up the high-frequency end of the response curve.

The design of most coupling arrangements is primarily based upon their response curves, from which there are high- and low-frequency limits which are the criterion points of acceptance. In other words, at frequencies above this high-frequency point and below this low-frequency point, the response is not acceptable to design purposes. These limitations are called the "half-power points" or "3 db points," and the high and low frequencies, upon which
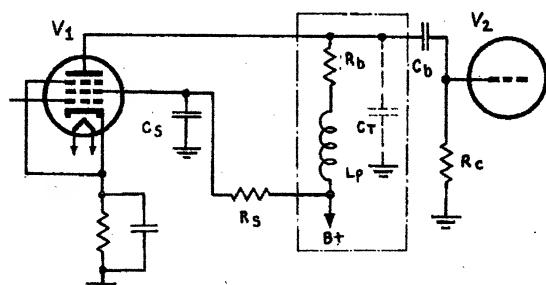

Fig. 27.—Coil L_p in this R-C coupled stage is inserted in series with the load resistor to increase the high-frequency response.

these limitations are based, are the determining factors upon which the design of certain coupling arrangements are founded. The half-power point or 3 db point is where the voltage amplification of the response curve has dropped to 70.7% of the maximum value, which occurs at the middle-frequency range of the band of frequencies in question.

For instance if a certain radio-frequency R-C coupling arrangement without a peaking coil has a response curve that is considered flat up to the half-power point of 1200 kc, then as far as design purposes go that R-C arrangement is only considered to be good up to the 1200-kc frequency region. However, if frequencies as high as 1600 kc are desired to be passed, then something must be done to the R-C coupling arrangement to increase the response at the high-frequency end and make the curve such that the 1600-kc point will be at *least* at the half-power point. Since the response curve has to be corrected to a degree where 1600 kc will be considered acceptable, then this 1600-kc frequency is termed the "frequency of correction." It is this frequency that aids in choosing the value of compensating components used in coupling arrangements. This will be seen when choosing the value for the peaking coil is analyzed. In order to have the voltage amplification at the half-power point equal to 70.7% of that maximum voltage at the middle frequencies, then the resistive load R_b on the tube must equal the reactance of the shunting capacitances X_{CT}.

This can be shown by examination of equation (1). If the resistance R_b be made equal to the reactance of the shunting capacities, X_{CT}, then equation (1) would be:

$$\frac{A_{high}}{A_{mid}} = \frac{1}{\sqrt{1 + \left(\frac{R_b}{X_{CT}}\right)^2}}$$

and if $R_b = X_{CT}$, then

$$\frac{A_{high}}{A_{mid}} = \frac{1}{\sqrt{1 + (1)^2}} = \frac{1}{\sqrt{2}}$$

$$= \frac{1}{1.414} = .707 \quad (3)$$

From equation (3) it is seen that when R_b is equal to X_{CT}, then the voltage amplification

at the high frequencies has fallen to .707 (70.7%) of that voltage amplification at the middle frequencies. This is the 3 db or half-power point mentioned before.

The choice of a load resistor depends upon the amount of shunting capacitances; therefore,

$$R_b = X_{CT} = \frac{1}{2\pi f_c C_T} \quad (4A)$$

where f_c equals the frequency at which correction starts. If the above relation is solved for $2\pi f_c$, it will be found that

$$2\pi f_c = \frac{1}{R_b C_T} \quad (4B)$$

Since at resonance the inductive reactance equals the capacitive reactance, then the inductive reactance of L_p is likewise equal to R_b. This may be expressed as follows:

$$R_b = 2\pi f_c L_p \quad (5)$$

since $2\pi f_c = \frac{1}{R_b C_T}$, it can be substituted in equation (5) from which we have

$$R_b = \frac{L_p}{C_T R_b} \quad \text{or rearranging,}$$

$$L_p = C_T R_b^2 \quad (6)$$

Equation (6) is a direct formula which shows the amount of shunt peaking coil inductance to use when both C_T and R_b are known. Notice that this latter relation is independent of frequency. It has been found from practice that when a response curve is plotted for all values of frequency versus gain (or voltage) using the compensated network of Fig. 27 and using a value of L_p as determined by equation (6), the high-frequency end of the curve is improved considerably over the ordinary R-C coupled circuit. The amplitude or gain of the response curve at the high-frequency end receives the improvement but the *overall* flatness of the curve is not improved as would be desired. The reason for this is because the value of L_p used, as calculated by $C_T R_b$, is too large a value and has caused *too* high a rise in amplitude at the high end of the curve. It immediately becomes apparent why L_p is called a "peaking" coil. From actual experimentation it has been found that for good overall flat response, down to the

frequency where correction is desired, the peaking coil should have a value equal to about *one half* of that as calculated by the above equation. In other words, the peaking coil L_p should be as follows:

$$L_p = \tfrac{1}{2} C_T R_b{}^2 \qquad (7)$$

The above relation is the one generally accepted as the shunt-peaking coil design formula. However there are many cases where the total shunt capacities C_T are not known accurately and it will then be difficult to find what L_p should be.

However, it is known that $R_b = \dfrac{1}{2\pi f_c C_T}$ and if this relation is solved for C_T, we find that $C_T = \dfrac{1}{2\pi f_c R_b}$. If this latter relation for C_T is substituted in equation (7) then,

$$L_p = \tfrac{1}{2} \times \frac{1}{2\pi f_c R_b} \times R_b{}^2 = \frac{R_b}{4\pi f_c} \qquad (8)$$

In this equation L_p is entirely independent of the shunt capacitances C_T and is dependent upon the frequency of correction f_c and the load resistance R_b.

Up to this point it should be remembered that even though one equation may be independent of certain component parts and dependent upon others, one equation cannot exist without another. For instance, the primary purpose of this type of compensation is to get rid of the effect of C_T and all computations have to be based on the value of C_T. In fact, R_b, for correction purposes of this circuit, has to be equal to the reactance of one-half of C_T at the frequency of correction. However, if R_b is already known by approximating a certain value for C_T, which is often done, then the above formula for L_p will hold.

As an illustration of how different values of shunt-peaking inductances affect the high-frequency end of the response curve, let us examine some typical curves at the high-frequency end. In Fig. 28 are a number of different curves at the high frequencies, in which each curve is representative of a different value of L_p. Note that the curve where the response is flattest is when

$$L_p = \tfrac{1}{2} C_T R_b{}^2$$

In order to appreciate the use of the preceding formulas let us apply them to arbitrary circuits where certain component values are known. For instance, in an R-C coupled network between two r-f stages it has been estimated from the response curve for this network, that the total shunting capacitances existing across the load on the amplifier are approximately 25 mmf. The circuit is used in the broadcast band and it has been found that the response curve falls off at about 1400 kc
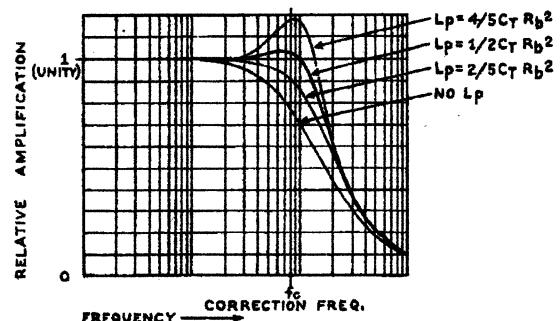
Fig. 28.—Typical curves showing the high-frequency correction due to the insertion of L_p in the amplifier plate load.

and it is desired to have good response up to about 1600 kc. In order to peak up the response curve at 1600 kc, it was decided to insert a shunt peaking coil. It has been found that for best overall flat response, the load resistance R_b should be equal to the reactance of the shunting capacitances at the frequency of correction. Mathematically it means:

$$R_b = X_{CT} = \frac{1}{2\pi f_c C_T}$$

which is the same as equation (4A). From this latter relation the value of the load resistor needed can be calculated. We know that $f_c = 1600$ kc and $C_T = 25$ mmf, then,

$$R_b = \frac{1}{2\pi f_c C_T}$$

$$= \frac{1}{2 \times 3.14 \times 1600 \times 10^3 \times 25 \times 10^{-12}}$$

$$= 3960 \text{ ohms}$$

Therefore we know that the load resistance should be equal approximately to 4000 ohms. From equation (7) or (8) the required value of L_p can be found.

From (7)

$$L_p = \tfrac{1}{2} C_T R_b^2$$
$$= \tfrac{1}{2} \times 25 \times 10^{-12} \times (4000)^2$$
$$= \tfrac{1}{2} \times 25 \times 16 \times 10^{-6}$$
$$= 200 \times 10^{-6}$$
$$L_p = .2 \text{ millihenry}$$

Or from (8)

$$L_p = \frac{R_b}{4\pi f_c}$$

$$= \frac{4000}{4 \times 3.14 \times 1600 \times 10^3}$$

$$= \frac{4000}{20.1 \times 10^6}$$

$$= 199 \times 10^{-6}$$

$$= .199 \text{ millihenry}$$

which is close enough to be called .2 millihenry.

Series Peaking

There is another type of high-frequency compensation using an inductance as a peaking coil and it is known as "series peaking." In this type of compensation the coil is placed *in series* with the plate of the amplifier tube instead of in parallel with it. In Fig. 29A is shown a typical series-peaking compensating network. The series-peaking coil, designated as L_s, is inserted between the load resistance R_b and the coupling capacitor C_b. The primary purpose of inserting L_s in this position is to *isolate the shunting capacitances C_1 and C_2* so that each will function individually. The combination of C_1, L_s, and C_2 represents a "filter" network, which has the characteristics of a low-pass filter (wide-band). Such reference to

a filter circuit means one so designed that it will pass certain frequencies within a given range and attenuate others that are undesired. That is why series-peaking compensation is sometimes known as "filter" coupling.

There may appear in certain high-frequency compensating circuits, a few variations of Fig. 29A; some of these variations are shown in Figs. 29B and 29C. The principle difference between these circuits is in the manner of isolating the shunt capacitances C_1 and C_2. In other words, as will be seen later on, the ratio between C_1 and C_2 is the fundamental expression upon which the choice of a compensating coil is based. For instance in Fig. 29B, the series-peaking coil L_s is shifted to the other side of the load resistor R_b because in some circuits the shunting capacitances are more evenly distributed under these circumstances. Or, as illustrated in Fig. 29C, the peaking coil may be inserted between the coupling capacitor C_b and the grid circuit of tube V_2, because it is desired to add the shunting capacitance between capacitor C_b and ground to C_1 and not to C_2, as in Fig. 29A.

For most practical purposes it has been found from experimentation that C_2 should approximately be equal to twice C_1. In practice, C_1 is usually lower than C_2 and it is the custom to place the load resistor R_b on the side of the lower shunting capacitance, namely C_1, as in Fig. 29A. The reason for placing R_b in such a position is to have the value of this resistance based upon the value of C_1, the lower shunting capacitance. In other words, the value of the load resistance R_b depends upon the reactance of the shunting capacitances across it. This was similarly analyzed in the discussion on shunt peaking, where it was stated that for
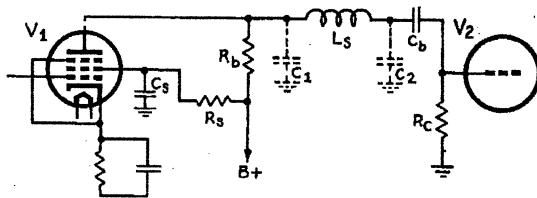

FIG. 29 A

FIG. 29 B

FIG. 29 C

Figs. 29A, B, and C.—These three schematics illustrate different methods of inserting a series peaking coil in an R-C coupled network.
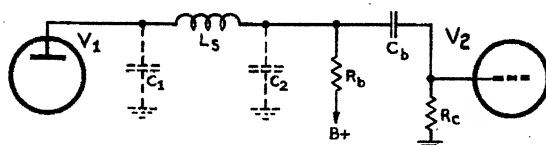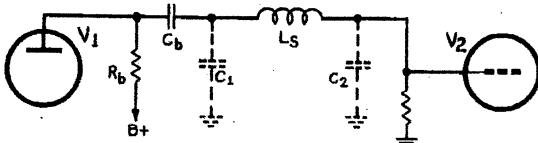
proper operation of R-C coupling the design basis of the load resistor is that it should be made approximately equal to the reactance of the shunting capacitances existing across the resistance. Consequently, the lower C_1 will be, the higher its reactance will be, and the higher resistor R_b will have to be. Thus it is seen that by isolating C_1 and C_2 properly, the effective load on the tube can increase, thereby allowing for increased gain. However, the effect of reducing C_1 is only part of how L_s affects this network.

The way the addition of the peaking coil L_s helps increase the high-frequency response is fully explained as follows: The voltage output from tube V_1 is impressed across the plate-load resistor R_b. Assuming no compensation (L_s removed), then this voltage across R_b would be attenuated at the higher frequencies due to the shunting effect of the total shunting capacitance. With peaking coil L_s inserted, the total shunting capacitances are divided, with the smaller amount appearing as C_1, as far as design features will allow. Therefore, the previous loss, which was due to the effect of the total shunting capacitances, is greatly reduced because it is now due to C_1 alone. However, C_1, even though reduced, still causes somewhat of a loss at the high frequencies. The output voltage from tube V_1, that is now developed across R_b and C_1 in parallel, is impressed onto the voltage dividing network of L_s and C_2. Capacitor C_b presents a virtual short circuit at high frequencies and the resistance of the grid-leak R_c is so large compared to the reactances of L_s or C_2 that C_b and R_c can be considered as having negligible effect on the voltage dividing network of L_s and C_2, and are consequently neglected in the circuit analysis at high frequencies. Even though the shunting capacitance across tube V_1 is now reduced from the total shunting capacitance C_T to just the shunting capacitance of C_1, there still remains some loss at high frequencies due to this C_1. The problem now is to offset this high-frequency attenuation caused by C_1. This can be done by producing a resonant rise in voltage across C_2 at high frequencies. The question that immediately arises is, how can this effect be attained?

Referring to Fig. 30, which is a simplified circuit of Fig. 29A, it is found that at some frequency, f_c, L_s and C_2 will be in series resonance. Therefore, at this frequency the volt-

age appearing across C_2 will be greater than the voltage impressed across the whole resonant circuit (L_s and C_2) by the preceding tube (V_1). (See the introduction to this section). The maximum amount of voltage (at high frequencies) will now be impressed on the grid of tube V_2 by the voltage drop across the capacitance C_2. Therefore, it is seen how capacitor C_2 offsets the drop in gain at high frequencies by the characteristics of its resonant effects, at these frequencies, with the series peaking inductance L_s. In actual practice, the ratio between C_2 and C_1 is set approximately equal to 2, as stated before. The value of C_1 can be measured accurately and so can the total capacitance C_T ($C_T = C_1 + C_2$), and from knowing C_1 and C_T, C_2 can be found. The physical location of the circuit components may have to be shifted in order to make the ratio correct, since C_1 and C_2 are dependent on stray capacitances to a large extent.

Both C_1 and C_T may be measured in the same way and the usual method of making this measurement is indirect. This method depends upon the relationship which exists between the load resistance and the stray capacitance at the high-frequency half-power point. As was stated previously, the meaning of the high-frequency half-power point is that at a certain relatively high frequency, the voltage gain of an amplifier decreases to 70.7% of the gain available in the middle range of the band of frequencies for which the amplifier is designed. At this half-power point, in an uncompensated amplifier, the load resistance equals the reactance of the total capacitance (C_T) shunting it.

In order to make the actual measurement of C_T, the amplifier to be compensated is set up in a form as close as possible to the final form. This is important because of the dependence of C_T (and C_1 alone) on stray capacitance, which
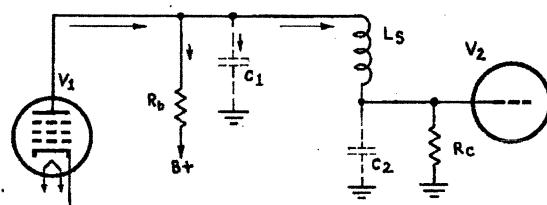

FIG. 30.—The schematic of Fig. 29A is here redrawn to show how L_s and C_2 are considered as a series resonant circuit. Capacitor C_b is removed, because at high frequencies it is considered as a short circuit.

in turn depends upon the circuit layout. For example, a peaking coil of the approximate size expected should be included in the layout. This coil is not connected when C_T is measured, but its physical presence is required because of its effect on stray capacitance. Since the value of the load resistor R_b in the completed amplifier depends upon C_T, it is possible to use only an approximately correct value of R_b during this measurement. This is not important, however, since C_T is affected only by the mechanical size of R_b, not by its electrical value.

When the amplifier has been set up, its voltage gain is measured over a range of frequencies, beginning somewhat below the middle of the expected band and extending beyond the highest frequency desired. A point of maximum gain will be found at the middle frequencies, and at the higher frequencies the gain will decrease. At the high frequency where the gain drops to 70.7% of the maximum, the reactance of C_T is equal to the resistance of R_b (the load resistor actually in use during the measurement). When the reactance of a capacitor (in this case C_T) at a certain frequency (the high-frequency half-power point) is known, it is an easy matter to calculate the unknown capacitance from the previous relation that R_b equals X_{C_T} at the half-power point. Thus C_T is measured indirectly.

Since C_i is also a distributed capacitance, it can be measured in much the same way. Since C_i is part of C_T ($C_i = C_T - C_s$) and if the undesired (for this measurement) part of C_T is removed from the circuit, C_i alone can be measured by understanding that R_b is equal to X_{C_i} at the half-power point. The undesired part (C_s) is removed by disconnecting and, if necessary, mechanically removing the components across which C_s appears (see Figs. 29A, B, C).

It has been shown that in the ordinary R-C coupling arrangement, the high-frequency end of the response curve falls off. It has been found that for series-peaking compensation with C_s equal to about twice C_i and the elements L_s and C_s made to resonate at the desired frequency of correction f_c, then the R-C coupling network is said to be effectively compensated. Series-peaking compensation gives a higher gain than shunt-peaking compensation because the effective shunt capacitances across R_b are reduced, thereby allowing for an increase in the load resistor R_b. And since the gain of an

R-C coupled stage is proportional to R_b and inversely proportional to the shunt capacitances, the above statement then holds true.

Up to this point the series-peaking circuit has been analyzed from the standpoint of why the coil is inserted and how it works. Some simple mathematics are needed for final clarification as to the representative methods of how the values of the series-peaking coil and load resistance are obtained. The following formulas, as those previously encountered, are not difficult. One simple formula leads to another by the easy method of substitution.

From the design procedure, it has been stated that at the highest frequency where correction is desired, the series circuit of L_sC_s should be made resonant at the frequency, f_c. Consequently, from the resonant-frequency formula, it is found that:

$$f_c = \frac{1}{2\pi\sqrt{L_sC_s}} \qquad (9)$$

If both sides of this equation are squared and solved for the inductance L_s, we have:

$$f_c{}^2 = \frac{1}{(2\pi)^2 L_sC_s} \quad \text{and rearranging}$$

$$L_s = \frac{1}{(2\pi f_c)^2 C_s} \quad \text{or} \quad 2\pi f_c L_s = \frac{1}{2\pi f_c C_s} \qquad (10)$$

The above equation (10) is dependent upon the shunting capacity C_s alone, and upon no other circuit components. If we want to solve for L_s in terms of C_i, then from the relation that $C_s = 2C_i$ we have:

$$L_s = \frac{1}{(2\pi f_c)^2 2C_i} \quad \text{or} \quad 2\pi f_c L_s = \frac{1}{2}\frac{1}{2\pi f_c C_i} \qquad (11)$$

Equation (10) states that the inductive reactance of L_s is equal to the capacitive reactance of C_s at the correction frequency f_c and equation (11) states that the inductive reactance of L_s is equal to one half of the capacitive reactance of C_i.

If equation (11) is attacked in a different way and solved for the relations $\sqrt{2}\,f_c$, we would have the following:

First, by taking the square root of (11) we have:

$$\sqrt{L_s} = \frac{1}{2\pi f_c\sqrt{2C_i}}$$

Then if this latter relation is solved for $\sqrt{2}\,f_c$ we have:

$$\sqrt{2}\,f_c = \frac{1}{2\pi\sqrt{L_s C_l}} \qquad (12)$$

(At this point it would be wise to understand something about square roots. As an illustration we will consider the expression $\sqrt{2C_1}$. This $\sqrt{2C_1}$ can also be represented as the $\sqrt{2} \times \sqrt{C}$. In other words, the product of any two square root numbers, $\sqrt{2} \times \sqrt{C}$, will give the same answer as the square root of the product of the numbers $\sqrt{2C}$. For instance, if we let C be equal to 10 mmf, then $\sqrt{2C}$ is equal to $\sqrt{2 \times 10} = \sqrt{20}$, which is equal to 4.47 mmf. Similarly, the $\sqrt{2} \times \sqrt{C}$ is equal to the $\sqrt{2} \times \sqrt{10}$. The $\sqrt{2}$ is equal to 1.414 and the $\sqrt{10}$ is equal to 3.160. When these values are multiplied together, 1.414 $\times$ 3.16, they are also equal to 4.47. Therefore, it can be seen that $\sqrt{2C} = \sqrt{2} \times \sqrt{C}$ and any sort of multiplication similar to this would also hold true.)

Returning to equation (12), it will be noticed that the right-hand side of the expression is itself representative of a resonant frequency. In other words, a frequency exists at which the series-peaking coil L_s is in resonance with the lower valued shunting capacity C_l. The resonant frequency of $L_s C_l$ is then equal to $\sqrt{2}$ times the frequency of correction. That is to say, if we let f_r equal the resonant frequency of L_s and C_l, then

$$f_r = \sqrt{2}\,f_c \qquad (13)$$

It has been found *from practice* that the plate-load resistance R_b had to be equal to about one half of the reactance of the shunting capacitance C_l at the frequency of correction. Therefore,

$$R_b = \frac{1}{2}\,X_{cl} = \frac{1}{2}\,\frac{1}{2\pi\,f_c\,C_l} \qquad (14)$$

where X_{cl} is equal to $\dfrac{1}{2\pi\,f_c\,C_l}$, the capacitive reactance at the frequency of correction of the shunting capacity C_l existing across R_b. Equations (11) and (14) are equal to the same thing so they are therefore equal to each other and we have the following relation:

$$2\pi\,f_c\,L_s = R_b \qquad (15A)$$

which states that the inductive reactance of L_s is equal to the load resistance R_b. This may be rewritten as:

$$L_s = \frac{R_b}{2\pi\,f_c} \qquad (15B)$$

Equation (15B) can be used to determine the value of series-peaking inductance when the value of the load resistance and frequency of correction are known.

All of the above equations, (9) to (15), have been analyzed for the case where $\dfrac{C_s}{C_l} = 2$.

However, as previously mentioned, this ratio is not always easy to attain in practice and in other cases, this ratio may not be desired. For this reason formulas for R_b and L_s have been derived to cover any ratio between C_s and C_l. Calling this ratio "S", then from equation (14) we have:

$$R_b = \frac{1}{\sqrt{2S}\;2\pi\,f_c\,C_l} \quad \text{or} \quad \frac{1}{\sqrt{2S}}\,X_{cl} \qquad (16)$$

And from equation (11) we have:

$$L_s = \frac{1}{\sqrt{2S}\;(2\pi\,f_c)^2\,C_l} \quad \text{or} \quad \frac{1}{\sqrt{2S}\;2\pi\,f_c}\,X_{cl} \qquad (17)$$

If the shunting capacitances across the load resistor in a series-peaking network were known and the frequency of correction known, then the above formulas could be used to determine the necessary value of R_b and L_s. Let us consider the following examples.

It was found that the response of an R-C coupled r-f amplifier dropped at the high-frequency end of the curve in the 540-to-1700-kc range of the broadcast band. Consequently, compensation at a correction frequency of 1700 kc was desired to increase the band-width characteristics. It was found that the shunting capacitances existing across the load resistance, with a series-peaking coil inserted as in Fig. 29A, was equal to about 15 mmf and that the ratio between C_l and C_s was about 2. Under these circumstances the value of the load resistor can be determined by use of equation (14). Thus, with $C_l = 15$ mmf,

$$R_b = \frac{1}{2} \times X_{c1} = \frac{1}{2} \quad \frac{1}{2\pi f_c C_1}$$

$$R_b =$$

$$\frac{1}{2} \quad \frac{1}{2 \times 3.14 \times 1700 \times 10^3 \times 15 \times 10^{-12}}$$

$$R_b = \frac{1}{2} \quad \frac{1}{1.605 \times 10^{-4}}$$

$$R_b = 3120 \text{ ohms}$$

The value of the series-peaking coil can be obtained from equation (15B). Therefore,

$$L_s = \frac{R_b}{2\pi f_c}$$

$$L_s = \frac{3120}{2 \times 3.14 \times 1700 \times 10^3}$$

$$L_s = \frac{3120}{1.07 \times 10^7}$$

$$L_s = 2920 \times 10^{-7}$$

$$L_s = 0.292 \text{ millihenry}$$

From the above analysis it can be stated that the load resistor needed is about 3100 ohms and the value of series-peaking coil about 0.29 millihenry.

Some fundamental properties of series-peaking compensation compared to shunt peaking follow. The former type of compensation gives approximately 50% more gain to the circuit than shunt peaking, where both circuits are compared under the same values of R_b, f_c and C_T. Series-peaking compensating networks offer a more linear phase shift to the signal being coupled from one stage to another.

So far two separate high-frequency compensating networks, namely shunt peaking and series

Fig. 31.—This circuit illustrates how series peaking and shunt peaking compensation are combined in one coupling circuit to give an overall response that is better than either one alone.

peaking, have been considered. The latter type of peaking was found to give a better response and phase shift than the former, but a combination of both of these types can give a better response than either one alone. Shown in Fig. 31 is a typical circuit combining the effects of series and shunt peaking with the shunting capacitances C_1 and C_2 distributed as shown. There is greater gain than if shunt or series peaking were used alone. It has been found that combination peaking can give as much as 75% more gain than shunt peaking alone (with $S = 2$, or $C_2 = 2C_1$). Actually, this means that the total *effective* load impedance is also 75% greater than that for shunt peaking.

Referring back to the two receivers mentioned at the beginning of this section, it is found that not enough information is obtained from these circuits to apply fully the formulas from the previous discussion. In other words, the values of the shunting capacitances are unknown or into exactly what ratio they are divided by their series peaking coils. All that can be done is to make assumptions of what the values may be, apply them to the formulas, and see how close the answers will be to the values of the components given by the manufacturer. It can be seen that these two circuits represent, in one form or another, those studied at the beginning of this section and consequently the analysis applies equally well here. The exact method that the designers of these models used in determining the values of load resistances and peaking coils is unknown, but doubtless they used methods similar to those previously described. That is to say, practically everything depended upon the knowledge of the shunting capacitances, and in what relation to them the load resistance and peaking coils were chosen. There are, however, a few interesting facts about these circuits that warrant recognition.

Let us first refer to the Espey Model RR-13L, as shown in Fig. 24. In this receiver the load resistance on the 7B7 r-f amplifier tube is equal to 3300 ohms. It is designated as R_1, which is the same as R_b used in the previous equations. The series peaking coil is designated as L_{sb}, which is the same as L_s used in the preceding discussion. The peaking coil is placed between the coupling capacitor C_s and signal grid circuit of the 14Q7 converter tube. This is similar to the circuit of Fig. 29C where the peaking
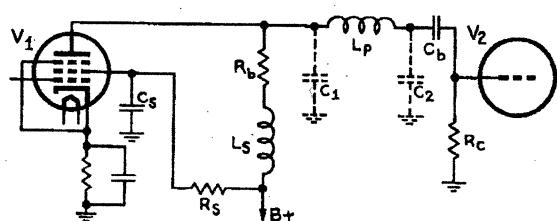
coil L_s is placed in the same position. The stray wiring capacitance existing between capacitor C_s and ground is therefore added to those shunting capacitances existing across the load resistor, R_1. The 6800-ohm resistor R_2 is used to "broaden" the peaking effect of the series peaking coil L_{sb}. It is explained as follows:

The peaking coil L_{sb} is in series resonance with those shunting capacitances existing across the signal-grid circuit of the 14Q7 tube. The resonant effects of the series network is such that the peak it produces at the high frequencies, is decidedly too great; in other words, the Q of the coil is too high. Consequently, a resistor, R_2, is inserted across this coil to reduce its Q and thus reduce the amplitude of the peak that occurs at resonance. The reduction in Q is enough to flatten the peak at the frequency of correction and make the response curve for the entire R-C compensated network level enough to give the desired characteristics. The output capacitance of the 7B7 r-f tube is about 7 mmf and the input capacitance of the 14Q7 tube is about 9 mmf (as found in the tube manuals.) These capacitances, added together with the stray wiring capacitances existing in the circuit around each tube, are the effective shunt capacitances of the circuit for which coil L_{sb} is compensating. However, the stray capacitances in the vicinity of the output of the 7B7 tube and those in the vicinity of the input to the 14Q7 tube, are difficult to estimate. They vary a great deal according to the method of wiring used. Such capacitances, in the vicinity of either tube, may vary anywhere from 5 mmf to 20 mmf. It is, therefore, obvious how difficult it is to apply the previous equations to this set, since they are all developed on the basis of knowing what the values, or the ratio, of the shunting capacitances are.

In the General Electric Model 321 (Early) (Fig. 25), a combination of series and shunt peaking is used. The same situation applies here as in the Espey receiver with respect to applying the previous equations to the actual components used in this model. The difficulty is again due to the fact that the values of the shunting capacities are unknown or at what frequencies the designer of the set desired the inductances to peak. If the values of the load

resistance and peaking coils were applied to the equations it would be found that they do not hold. Even though the exact values of C_1, C_2, or C_T are not known so as to apply them to the developed formulas, nevertheless, some conclusions can be drawn that will coincide with the introduction of this topic.

First of all, it is known that a certain amount of shunt capacitance exists across the output of the 12SK7 r-f amplifier and hence across the $R2$—$L2$ load circuit. These shunt capacitances will therefore be in parallel resonance with the inductance $L2$ at some frequency. The values of $R2$ and $L2$ are dependent upon the shunting capacitance across them and the frequency at which the circuit is designed to be resonant. A similar situation holds true for the series peaking coil $L4$. It is inserted in the circuit to isolate the shunting capacitances existing between the 12SK7 and 12SA7 tubes and also to be in series resonance with that shunting capacitance existing across the input to the 12SA7 tube.

The exact frequencies for which each coil is made to resonate is unknown. It is known, however, that such coils are inserted in the circuit to peak the response curve and broaden the band-pass characteristics of the R-C arrangement. Consequently, two facts can be deduced about this circuit, either of which may be correct. Inductances $L2$ and $L4$ may both be used to peak at the same frequency if the loss at some particular high frequency may be quite high. Or, the resonant frequency between both circuits may differ somewhat to the extent that their resonant peaks may be slightly apart. This will result in an increase in voltage gain at a wider range of high frequencies than if both coils were peaked at the same frequency.

The 270,000-ohm resistor across the series peaking coil $L4$ is used to broaden the resonant peak caused by the series peaking coil $L4$, similar to that in the Espey receiver. In this circuit, however, the resistance is much higher and the effective Q of the coil is not reduced as much as it was in the Espey model. From this fact it can be concluded that there apparently was a higher resonant peak caused by the series peaking coil in the Espey set than that caused by the one in the General Electric receiver.

MECHANICAL ARRANGEMENTS OF I. F. TRANSFORMERS IN RECENT RECEIVERS

In this section are given the exploded views of various representative i-f transformers used in some of the receivers, schematics of which are contained in Volume XV.

By combining the various capacitors and resistors in the can with the transformer, the manufacturer saves a considerable amount of space and avoids some production — line troubles. It would appear from this construction that the best way to service trouble in the i-f stages of a receiver using this type of coils would be to replace the entire assembly. The R-C network shown in most of the schematics of these coils is used for i-f rejection.

In showing the exploded views, we have in all cases depicted the mica insulating plates by a single line, and the metal plates by a double line. In the schematic drawings, the mica plates are shown by the conventional dotted lines.

The International Detrola Model 571X receiver incorporates the i-f rejection filter circuit components in the second i-f transformer can in an ingenious manner. The electrical schematic of this assembly is shown in Fig. 1A, an exploded mechanical drawing of the assembly is shown in Fig. 1B, and the composite mechanical assembly is shown in Fig. 1C.

The primary and secondary trimmer capacitors are conventional compression mica units. As shown in Fig. 1B, the adjustment screws of the trimmers make use of the metal washers to press the slightly bent spring capacitor plates together (A with C and B with D). The mica plates between the washers and the capacitor plates, serve to insulate the screws from the capacitor plates. The mica plates between the capacitor plates are the capacitor dielectric, while the mica plates between the capacitor plates and the base of the unit insulate the capacitors from the brass nuts in the base in which the adjustment screw fits. The i-f rejection filter fixed capacitors are circular metal and mica plates, which are mounted on the hub of the i-f transformer coil-form as-
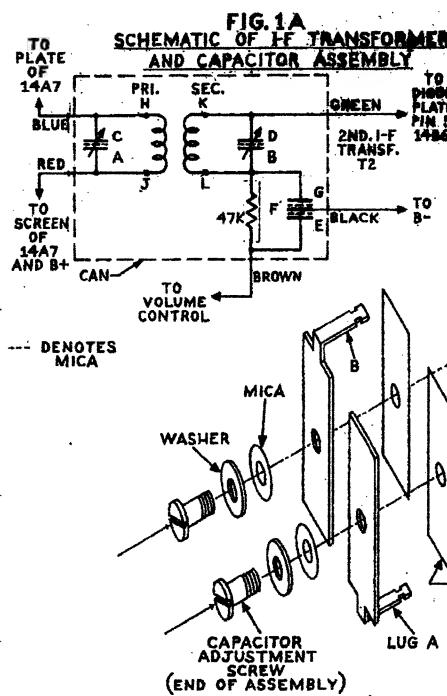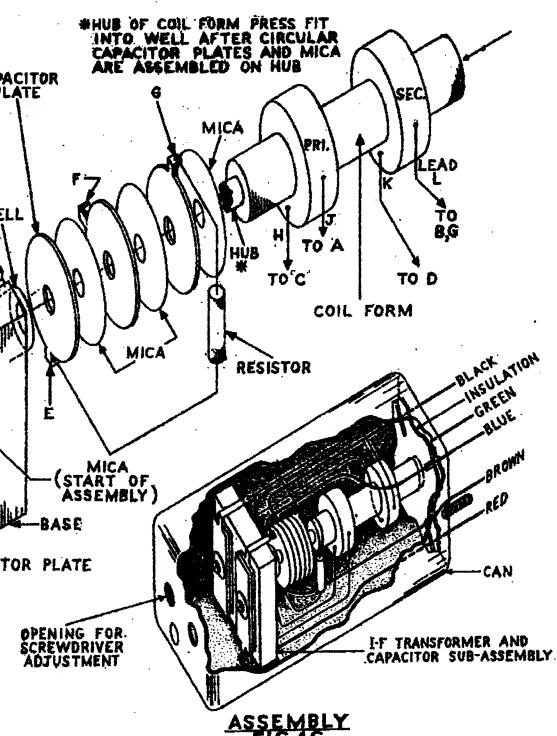

Fig. 1.—The components of the i-f rejection circuit are incorporated in the second i-f transformer can in the International Detrola Model 571X receiver.

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SCHEMATIC OF I-F TRANSFORMER AND CAPACITOR ASSEMBLY
FIG. 2A

EXPLODED VIEW OF I-F TRANSFORMER AND CAPACITOR SUB-ASSEMBLY
FIG. 2B
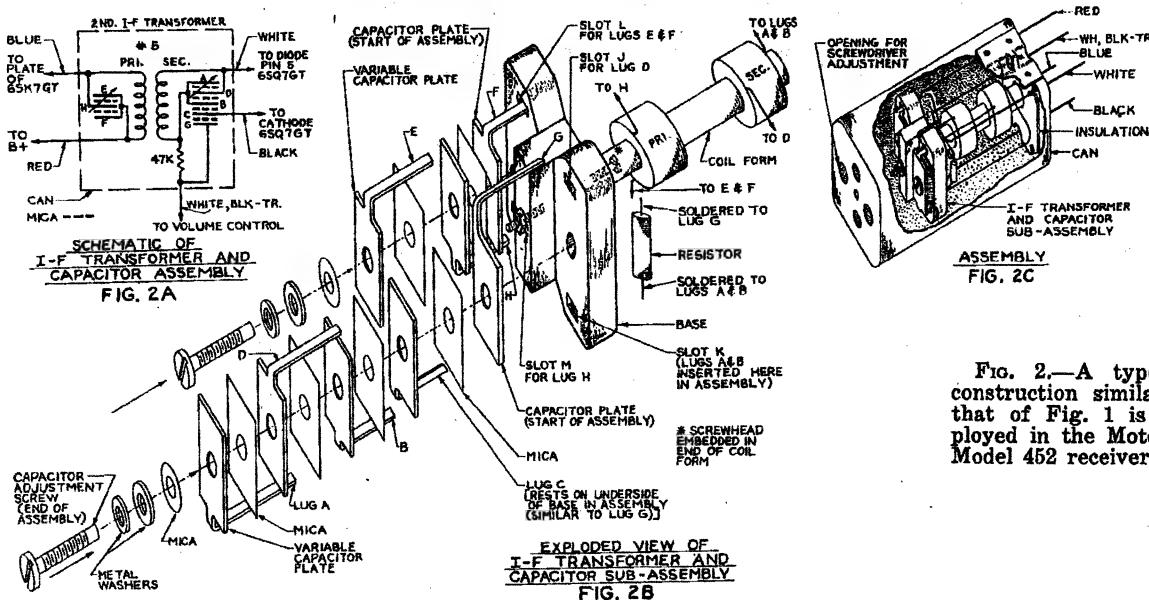
ASSEMBLY
FIG. 2C

FIG. 2.—A type of construction similar to that of Fig. 1 is employed in the Motorola Model 452 receiver.

sembly, the hub end then being pushed into a hole in the base so as to make a compact assembly. The filter capacitors make use of a common plate F connected to B minus, so that one capacitor plate is eliminated. The 47,000-ohm filter resistor connected between lugs G and E, is mounted entirely within the i-f can.

The assembly of the i-f transformer and i-f rejection filter circuit in this manner permits the elimination of two separate mica capacitors by the use of the compact mica and metal plates. This is less expensive, facilitates receiver assembly, saves space, and permits the use of very short leads.

The Galvin Motorola receiver Model 452 uses a similar ingenious method of saving space. Fig. 2A shows the electrical schematic of the second i-f transformer and i-f rejection filter assembly incorporated in that set. The primary and secondary coils are each shunted by a fixed capacitor, H-F and D-B respectively, and a trimmer capacitor, H-E and D-A respectively. The purpose of using both a fixed and trimmer capacitor is to obtain the required circuit capacitance which is not achieved by the trimmer alone, and to obtain the required trimming action for which it is not necessary to vary the entire shunt capacitance. The primary and secondary capacitors save one capacitor plate each by making use of common plates $(H$ and $D)$ as a plate of both the fixed and trimmer capacitors.

The i-f rejection filter is conventional, consisting of two low-capacitance fixed capacitors, G-C and C-B, separated by a 47,000-ohm resistor. The arrangement is so designed that two capacitor plates are saved by using plate C for both capacitors, and using plate B for one of the filter capacitors and for the secondary fixed shunt i-f capacitor. Fig. 2B, an exploded mechanical drawing of the assembly, and Fig. 2C, the assembled unit, illustrate the compactness and simple design incorporated in the assembly. It was found when one washer is used under the adjustment screw head, the turning of the screw caused the washer to turn when the screw is tight, thus wearing the mica insulator. The use of two washers under the screw head as shown, prevents this wearing of the mica.

Somewhat indicative of the fact that this method of saving space is becoming more widespread can be seen in that the Bendix Model 646A receiver and the RCA Model 56X receiver incorporate almost similar electrical circuits for the i-f transformer and i-f rejection filter as the above Galvin model and employs generally the same mechanical assembly of the circuit components. Figs. 3A and 4A, the electrical schematics of these circuits in the Bendix Model 646A receiver, and the RCA Model 56X receiver respectively, and Fig. 3B and Fig. 4B, the mechanical drawings of these assemblies and Figs. 3C and 4C, the composite assemblies,
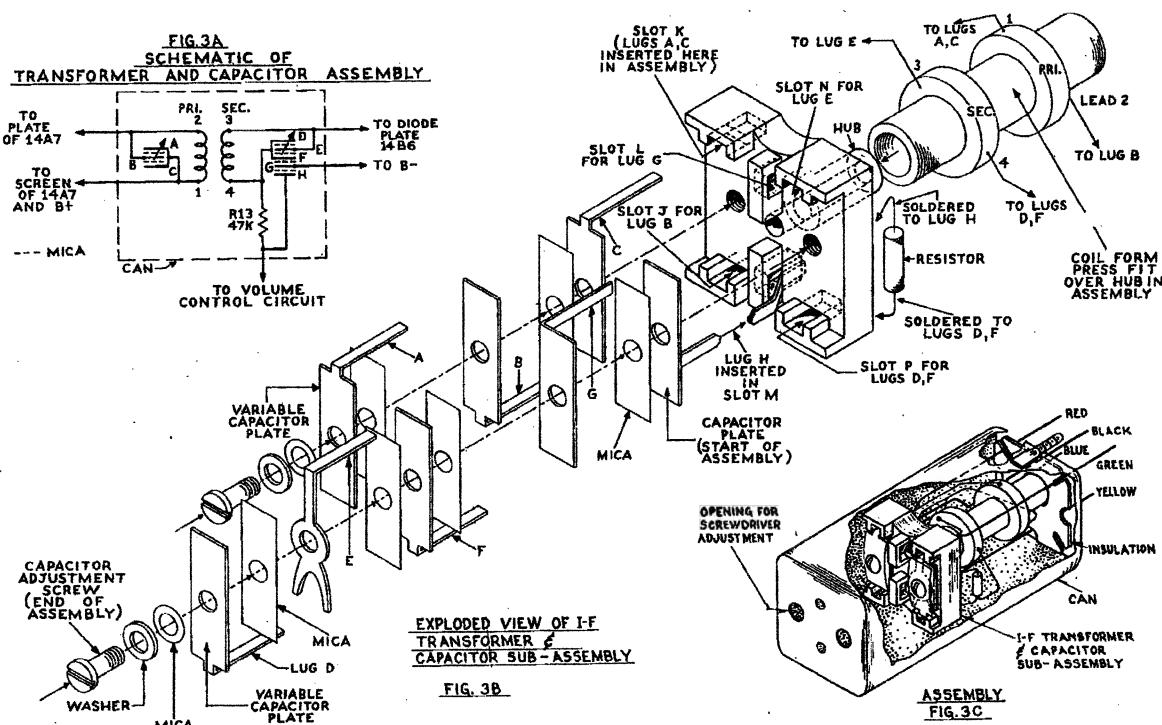
Fig. 3.—I-f rejection filter and i-f transformer as used in the Bendix Model 646A receiver.
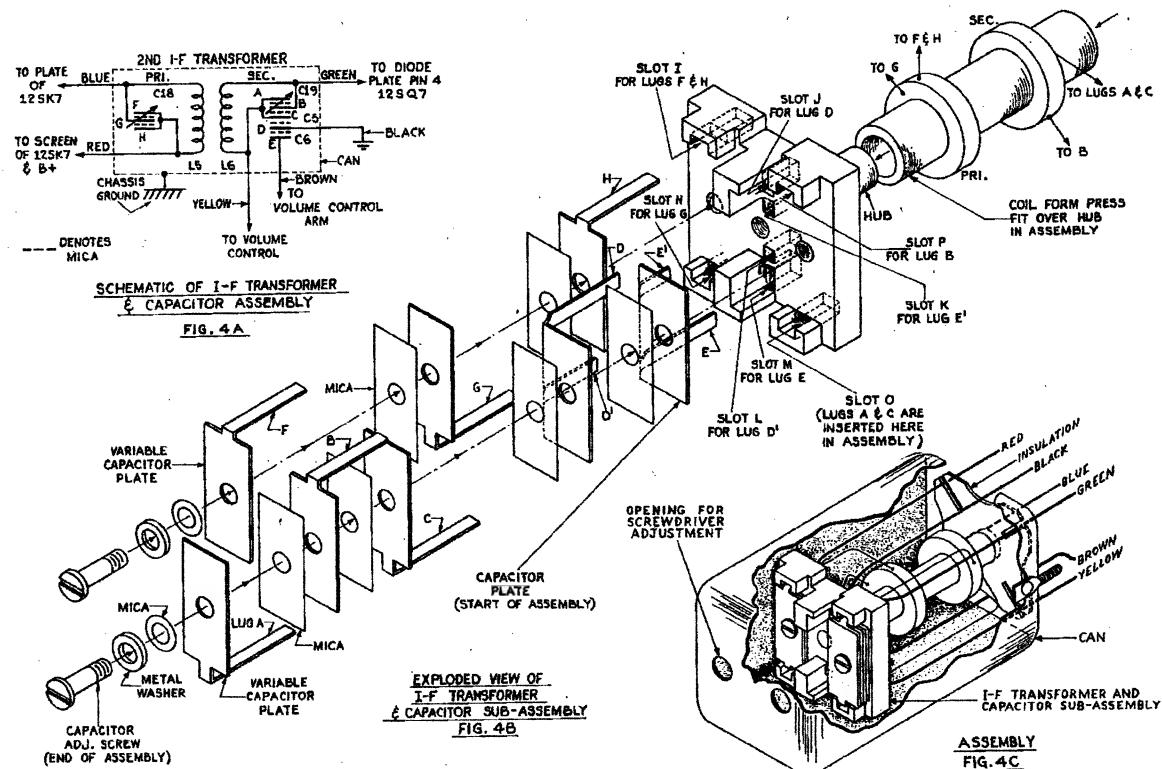
Fig. 4.—The RCA Model 56X receiver employs a similar i-f transformer construction as shown in the preceding illustrations.

do not differ markedly from the respective drawings (Figs. 2A, 2B and 2C) for the Galvin Model 452.

It should be noted, however, that changing the shape of any of the capacitor plates will change the value of the capacitor of which that plate is a part. Fig. 3B shows how the capacitor plate E had to be shaped, that is cut out, in order to obtain a smaller capacity, between capacitors D-E and F-E, which are in shunt with the secondary transformer coil.

The Zenith Model 5B042 receiver exhibits even more sharply the advantages of assembling the i-f filter circuit with the i-f transformer. Fig. 5C, the electrical schematic of this assembly, shows the transformer primary and secondary coils being shunted by trimmer and fixed capacitors, as in the receiver models

above. The i-f filter circuit consists of the volume control and its three shunt capacitors. These filter capacitors, however, are made up of the three parallel capacitances of the metal and mica plates between and including plate C to plate F shown in Fig. 5A, the exploded view and Fig. 5C, the schematic of the i-f transformer. These parallel capacitances are required for bypassing the i-f signal to ground. An equivalent simplified circuit showing this capacitor assembly is illustrated in Fig. 5D. Capacitances C-D, E-D and E-F from the transformer secondary low side to ground, shunt the volume control, whereas B-C and B-A are the fixed and trimmer capacitors shunting the secondary coil of $T2$. The mechanical drawing of the entire transformer and capacitor assembly is shown in Fig. 5E.

EXPLODED VIEW OF
I-F TRANSFORMER AND
CAPACITOR SUB-ASSEMBLY
FIG. 5A
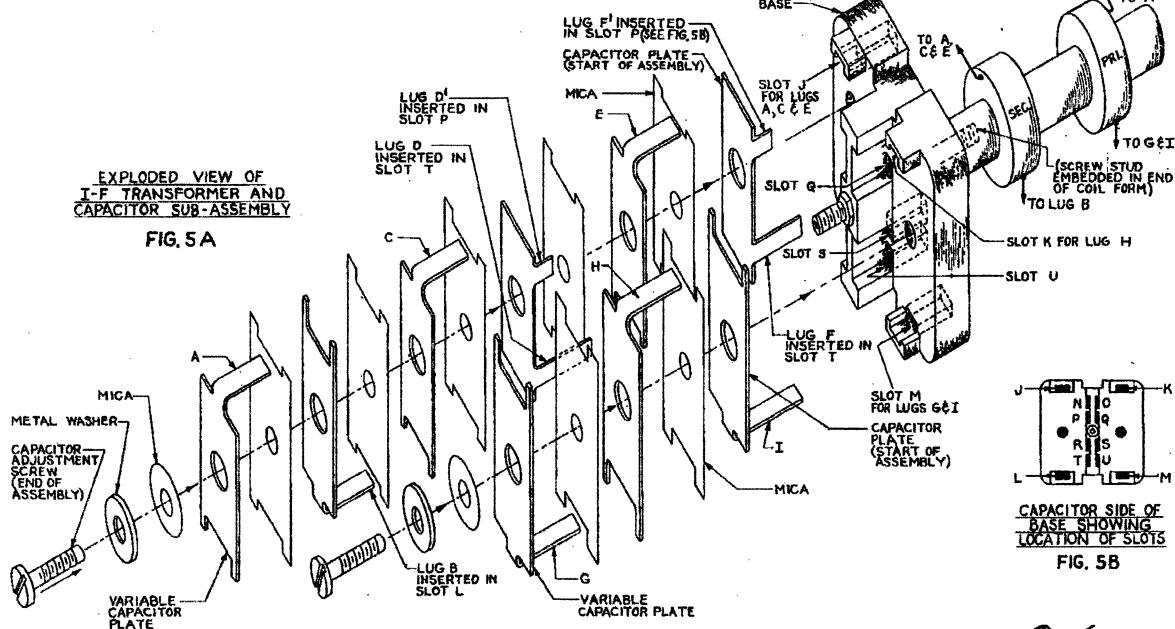
CAPACITOR SIDE OF
BASE SHOWING
LOCATION OF SLOTS
FIG. 5B
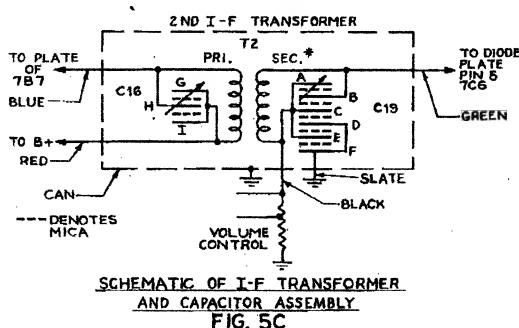
SCHEMATIC OF I-F TRANSFORMER
AND CAPACITOR ASSEMBLY
FIG. 5C
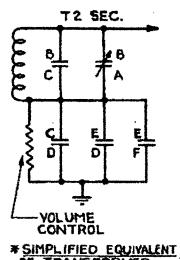
*SIMPLIFIED EQUIVALENT
OF TRANSFORMER
SECONDARY
FIG. 5D
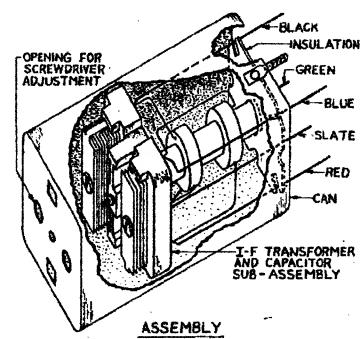
ASSEMBLY
FIG. 5E

FIG. 5.—The i-f filter circuit in the Zenith Model 5B042 receiver consists of the volume control and three shunt capacitors, which are built as part of the trimmer across the secondary of the i-f transformer, as shown in the above schematics.

The Zenith Model 12H090 AM-FM receiver is another example of space-saving methods. Besides the fairly old and common method of dispensing with different r-f and i-f tubes for AM and FM reception, this receiver makes use of two devices to save a great deal of space. The first device, shown schematically in Fig. 6A, consists of the two AM and two FM coils with their four tuning slugs of the first i-f transformer, enclosed in the same i-f can. The second device is the one previously described of making use of thin mica and metal plates assembled in the coil form base. Fig. 6B shows the method of mounting these coils and their associated capacitors within the can.

The sixteen components from the metal bottom plate to the nut, when assembled, comprise four fixed capacitors in use. The two double capacitors, a detailed drawing of which is shown in "View A", make up the four capacitors. Lugs 5, 7, 8, 1, 4 and 3 serve as the connection points to the capacitors. Lug 6

serves as the grounding point for the i-f transformer can, and lug 2 is not used in this particular assembly. Note that the contact portion of lugs 3 and 7 are very small in area. This portion of the lug fits directly on the metal plate C (View A) and thus forms a small capacitance between lugs 3 and 1 and lugs 7 and 5 for the FM coils; whereas a larger capacity exists between lugs 5 and 8 and lugs 1 and 4 for the AM coils.

Since this i-f transformer can thus encloses all the first i-f AM and FM coils, slugs, and shunt capacitors, and since the second, third, and fourth i-f transformer cans are similarly assembled, a considerable amount of space is saved. It is of interest to note on the Zenith Model 12H090 schematic (Zenith page 15-91), that the second, third, and fourth i-f transformers include five capacitors, which are assembled in a manner similar to that described for the first i-f transformer.

EXPLODED VIEW OF TRANSFORMER & CAPACITOR SUB-ASSEMBLY
FIG. 6B

FIG.6A—SCHEMATIC WIRING OF TRANSFORMER & CAPACITOR ASSEMBLY
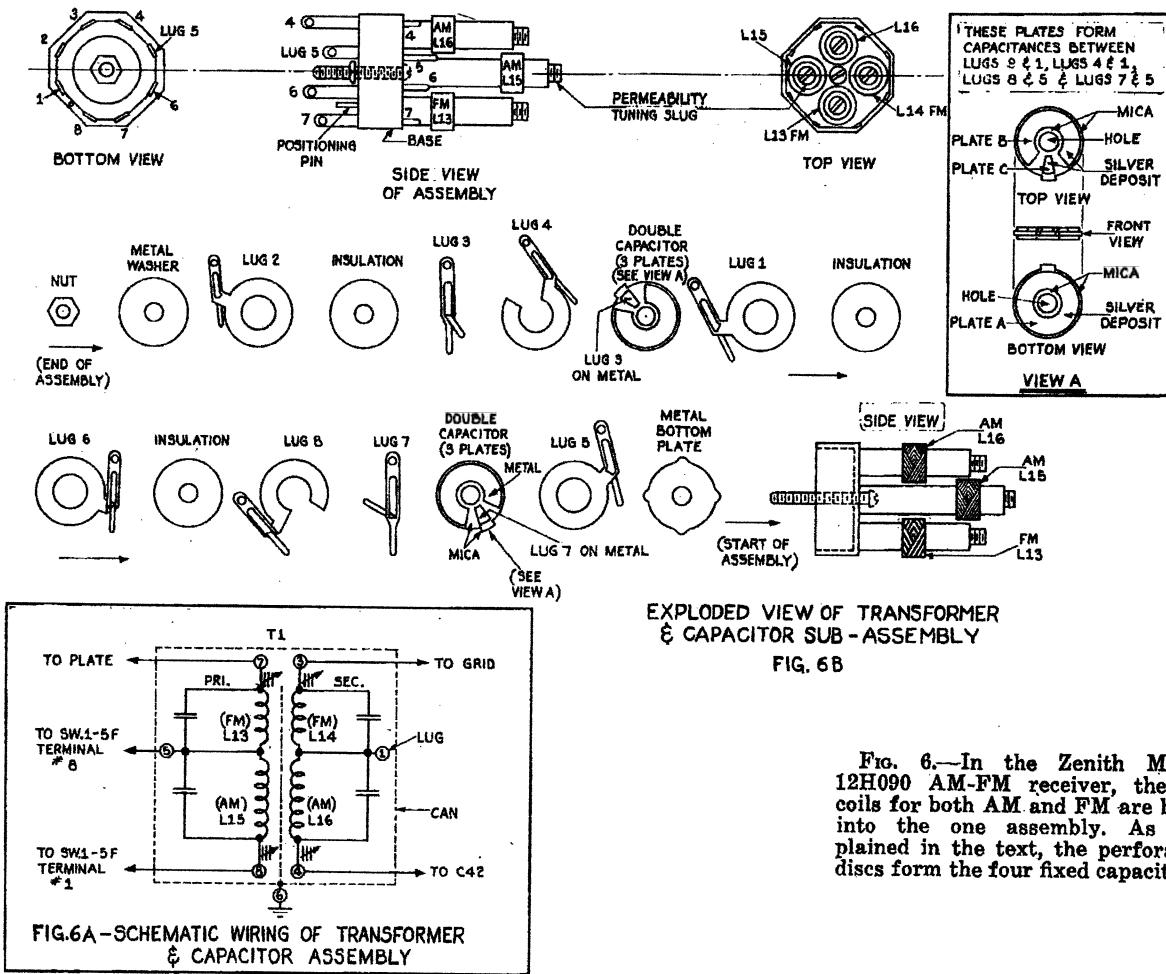
FIG. 6.—In the Zenith Model 12H090 AM-FM receiver, the i-f coils for both AM and FM are built into the one assembly. As explained in the text, the perforated discs form the four fixed capacitors.

FEEDBACK IN AMPLIFIERS

When a portion of a signal—whether it is r-f, i-f, or a-f—is fed back from the *output* of one stage to the *input* of the same stage or any preceding stage, then it is said to be a *feedback signal*. In most cases the receiver is designed for feedback and special circuits are inserted in the set to accomplish the effects of feedback. There are, however, different types of feedback that are inherent qualities of certain circuits and in most instances these kinds of feedbacks are undesired.

Feedback can be either of two forms. It may *aid* the original input signal (that is, increase the effective input signal) at the stage to which it is being fed back, or it may *oppose* it (that is, decrease the effective input signal). A number of standard terms exist that are used to express whether or not the feedback signal is aiding or opposing the original input signal. Such terms as *regenerative feedback, regeneration,* and *positive feedback* are used to indicate a signal that is fed back so as to *aid* the original input signal. On the other hand, such terms as *degenerative feedback, degeneration, negative feedback* and *inverse feedback* are used to indicate a signal that is fed back so as to. *oppose* the original input signal. Whenever any of these terms are used it should be understood what type of feedback is being considered.

When reference is made to audio, r-f, or i-f feedback signals, it is almost invariably accepted that the feedback signal is being returned to another stage of the same section of the set. That is to say, an audio feedback signal is returned to some part of the audio system of the set. Likewise, an r-f feedback signal is returned to some part of the r-f section of the set. Of course, a rectified audio signal is fed back from the audio stage to the r-f stage as an avc signal, but such a signal is considered as a d-c feedback voltage. Any reference to feedback in this section will be entirely based on a-c signals, either of an a-f, r-f, or i-f nature. For most general purposes when a feedback signal is regenerative, it means that this signal is *in phase* with the original input signal. On the other hand, when a feedback signal is degenerative, it means that this signal is *out of*

phase with the original input signal. At this point, let us consider a few facts relating to the term *"phase,"* so that a clearer knowledge of what is meant by "out of phase" and "in phase" will be known.

Phase and Phase Relationships

The term "phase" is used to denote a certain relation which exists between two or more periodic quantities of the same frequency. A periodic quantity is one that is alternating and has values recurring at equal time intervals. A sine wave is one of the simplest types of periodic quantities that can best be used to express phase relationships. Likewise since the type of periodic quantity used in the radio sets of today are sine waves or are composed of sine waves, then it follows that all references herein will be to such a wave.

In Fig. 1 is shown a typical sine wave. Note how the wave varies (alternates) above and below a certain reference level, called the *zero axis.* This wave is periodically alternating above and below the zero axis by 3 units; which units can be expressed either as voltage, current, or power. The time it takes to complete one alternation of the sine wave before it repeats itself periodically is called a *cycle* of the wave. Instead of representing time intervals along the wave by actual time in seconds or fractions thereof, the commonly accepted method is to express time in degrees. One cycle of a wave is completed in 360 degrees, one-half the cycle is completed in 180 degrees, etc., as shown in Fig. 1. When a cycle of the sine wave repeats itself, two cycles on the wave
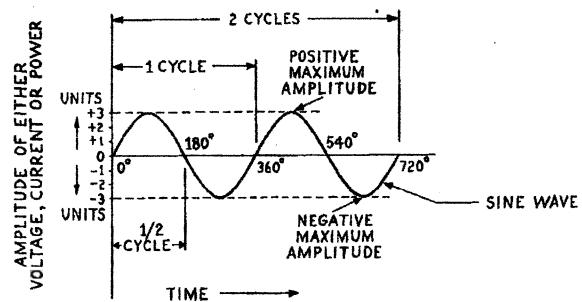

Fig. 1.—Two cycles of a typical sine wave. Note that time is indicated in degrees, one cycle of the wave being completed in 360°.

have occurred with the same amplitudes at corresponding points. In this latter case the "time" taken to complete the two cycles is expressed as 720° (degrees). If the sine wave was to repeat itself continually, each cycle would be adding another 360° to the time axis.

Since the second cycle is no more than a repetition of the first, then when time reference is made to a periodic function, such as the sine wave under discussion, it will be best to refer it as some value between 0 and 360°. This is so because the time difference between the advancing cycles of a sine wave are no more than an integer multiple of 360° and each cycle is exactly the same shape as the others. When there is reference to a frequency as 500 cycles per second, it means that for every second of time the sine curve undergoes 500 separate cycles. In other words, it takes 1/500 (or .002) of a second to complete one cycle. Since each individual cycle is completed in 360 degrees, regardless of its frequency, just one cycle can be considered as representative of a series of cycles; hence only the time between 0 and 360 degrees need be considered.

What would happen if two sine waves of the same frequency but of different amplitudes and starting at the *same time* were injected into a circuit at the same point? Since both waves start at exactly the same time, it means that when the positive half cycle of one wave travels through 0° to 180°, then the positive half cycle of the other wave also travels through 0° to 180°. The same can be said about their negative half cycles which vary between 180° to 360°. This is illustrated by a drawing as shown in Fig. 2.

It is readily seen how both sine waves Nos. 1 and 2 start at exactly the same time and pass through their maximum and minimum values at the same instant of time. Under such


Fig. 2.—Sine waves Nos. 1 and 2 have the same frequency but different amplitudes, and are in phase. Wave No. 3 is found by adding the amplitudes of Nos. 1 and 2, and so is greater than either of the two.

a circumstance, the curves will be additive, and the total input signal to the circuit in question will be greater than either one alone. This is illustrated by the heavy solid line, curve No. 3 in Fig. 2. From the above analysis it is seen how one signal can aid another signal and that together both signals offer a greater amplitude than either one alone.

If curve No. 2 is a signal voltage that is fed back from the output of a circuit to the input of another where the signal voltage of curve No. 1 exists, then the signal voltage of curve No. 2 is said to be *in phase* with that of curve No. 1. This is so because the signal voltage of curve No. 2 is aiding (i.e. additive in amplitude) to that of curve No. 1. In this respect the signal voltage (curve No. 2) fed back is of a *regenerative* nature with regard to that voltage already existing at the input source. Consequently, when a feedback voltage is regenerative it is said to be *in phase* with the original input signal.

If the fed-back signal voltage had a sine-wave characteristic that started its cyclic movement at a *different* time than that signal voltage existing at the input, then something different would happen. This is explained by the curves of Fig. 3. The same two sine waves Nos. 1 and 2 are used as the representative curves. Curve No. 1 is the sine wave existing at the input to a certain stage. Note that it starts its positive half cycle at the 0° reference point. Curve No. 2 is the sine wave of the feedback signal. The starting point of the positive half cycle of this latter curve is at point *A or 180° later than the starting point of curve No. 1.* However, at the same time curve No. 1 starts its positive half cycle, curve No. 2 starts its negative half cycle. Under these circumstances, when curve No. 1 reaches its positive maximum, curve No. 2 reaches its negative maximum; and likewise at the same instant curve No. 1 reaches its negative maximum, curve No. 2 reaches its positive maximum. Consequently, when both of these curves exist at the same point, their effects will be exactly opposite to that represented in Fig. 2. In other words, the resultant sine wave, which is the final effective input voltage, is such that its amplitude is less than the original input signal. In this respect the fed-back signal voltage is of a *degenerative* nature, since this latter signal *opposes* the original input signal. Therefore
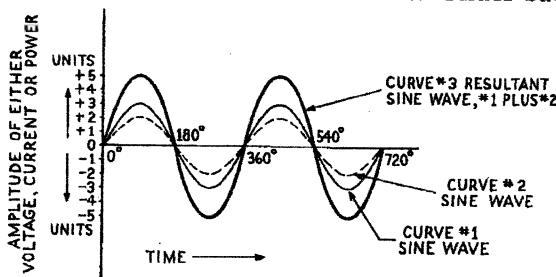
when a feedback signal is degenerative (as illustrated in Fig. 3), it is said to be *out of phase* with the original input signal. For most general purposes degenerative feedback in this section will be considered as being *180° out of phase with the original input signal*, as illustrated in Fig. 3. Since Fig. 3 represents pure subtractive sine curves, then a degenerative signal is also said to be *opposite in phase* to that signal already existing at the input circuit.

From the above analysis it can be concluded that such terms as *180° out of phase* and *opposite in phase*, describe the phase relationship
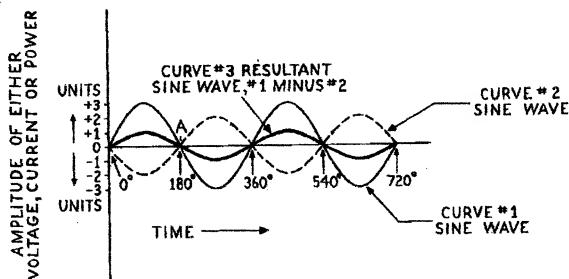

Fig. 3.—Sine waves Nos. 1 and 2 are 180° out of phase. Wave No. 3 is found by subtracting the amplitudes of the two waves and therefore has a smaller amplitude than either No. 1 or No. 2.

of degenerative feedback signals. Likewise, *in phase* represents the phase relationships of regenerative feedback signals.

Some Advantages and Disadvantages of Feedback

It was stated previously that sometimes feedback is desired and at other times it is not. From the preceding discussion it was shown that regenerative feedback *increases* the effective input signal to a stage and that degenerative feedback *decreases* the effective input signal to the stage. In either case there are definite advantages and disadvantages according to what part of the radio set the feedback occurs and, naturally, whether or not the feedback is desired.

Regenerative feedback is often used to increase the gain of a stage because the gain without feedback is not strong enough to operate the set satisfactorily. Regenerative feedback is also used in special circuits to increase only the bass notes and not the treble. There are also instances where regenerative feedback is an inherent quality of the set and the amount of feedback is so great that it produces oscillations in amplifier tubes, thus overdriving the

tube. This sometime occurs where a set is supplied by a common power supply. The power supply has a common impedance to all the tubes it supplies and if the filtering action of the power supply filter capacitors is unsatisfactory, then an audio signal may be coupled back to some stage through this common impedance of the power supply. Usually it is the output audio signal that is fed back to some audio amplifier stage, causing audio oscillations in the coupled stages and a resulting "singing" noise.

Degenerative feedback is used more often than regenerative feedback. The primary purpose of using degenerative feedback (very commonly called *inverse feedback*) in the different stages of the many radio sets, is to reduce distortion, reduce noise, and improve the flatness of the frequency-response characteristic. This type of feedback and the results it produces are readily evident when a cathode bypass capacitor is removed from across the cathode biasing resistor of an amplifier. For instance, if the cathode bypass capacitor was removed from an audio output stage the following would happen:

The signal currents that were originally bypassed around the cathode resistor by the capacitor will now flow through this resistor. By virtue of the intrinsic qualities of any tube, the signal voltage appearing on the cathode is in phase with that appearing on the signal grid of the same tube. This is because as the signal voltage on the grid goes in the positive direction, the current through the tube (the cathode current) increases. Conversely, as the signal voltage goes in the negative direction, the cathode current decreases. Now, when the cathode current increases, the voltage drop across the unbypassed cathode resistor also increases; and when the cathode current decreases, the cathode voltage will decrease. In a sense, then, although the average bias due to the cathode resistor may not vary when a signal is impressed on the grid, the bias at any one moment does change in step with the changes of the grid signal voltage. Consequently, the input signal voltage to the signal grid of the tube is *reduced* by the amount of signal voltage drop appearing across the cathode resistor. With this reduction in *effective* input voltage to the tube, the gain will be reduced.

In most general cases, however, the reduction in gain can be sacrificed especially if the tube has high gain characteristics. Along with the reduction in gain, there are other features which are very advantageous: these are reduction in distortion and noise, and a flatter frequency response. Also accompanying these desired features, degenerative feedback allows for greater stability to the tube with changes in operating voltages. Sometimes the removal of a cathode bypass capacitor does not introduce enough degenerative feedback and it must be obtained by different methods. There are numerous ways of obtaining degenerative feedback, as well as regenerative feedback, and some of the more interesting found in the schematics appearing in Rider's Volume XV, will be discussed later on.

To understand fully the topic of feedback it will be best to study it from a generalized viewpoint instead of any one particular type. From this aspect such related topics as what controls the amount of feedback, how the gain of the stage is affected, and a few others will be more clearly understood. Fig. 4 represents a block diagram of any amplifying feedback circuit where E_I is the original input signal. The amplifying circuit has a total gain equal to the letter A. The output signal from this amplifying circuit is designated as E_o.

In order that there be feedback from the output circuit to the input circuit, some of the signal output has to be *coupled back* to the input circuit. The sort of coupling arrangement is of no consequence at the moment, but what is important now is to realize that some type of coupling arrangement *has* to be used to feed back a portion of the output signal to the input circuit. In other words, the feedback
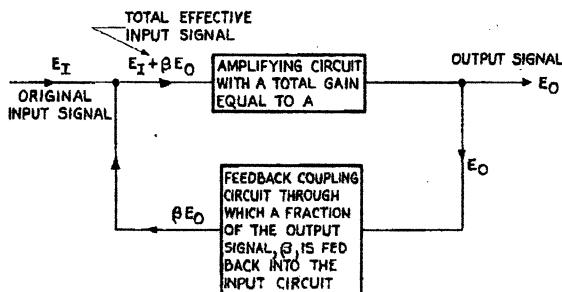
Fig. 4.—Block diagram of any feedback circuit which illustrates that a fraction β of the output signal E_o is combined with the input signal E_I to form the effective input signal.

coupling arrangement takes only a fraction of the output signal and returns it to the input circuit. This fraction is designated by the Greek letter β (beta) and the portion of the output signal fed back is designated βE_o.

The gain of any *system* whatsoever is given by the ratio of the output voltage divided by the original input voltage. In order to compute the gain of such a stage as in Fig 4, a few points should be understood. The *total effective* input voltage is the original input voltage E_I plus the feedback signal βE_o. (Whether the signal fed back is of a degenerative or regenerative nature will soon be seen). The output voltage E_o of the system is equal to the gain of the stage A multiplied by the total effective input signal. Consequently we have, mathematically represented, the following:

$$E_o = A \ (E_I + \beta E_o) \qquad (1)$$

Multiplying equation (1) out, it is found that:

$$E_o = AE_I + A\beta E_o \qquad (2)$$

Rearranging equation (2),

$$E_o - A\beta E_o = AE_I \qquad (3)$$

Factoring E_o out of equation (3),

$$E_o \ (1 - A\beta) = AE_I \qquad (4)$$

Solving equation (4) for E_o, we have

$$E_o = \frac{AE_I}{1 - A\beta} \qquad (5)$$

If both sides of an equation are either multiplied or divided by the same quantity, the significance of the equation will remain unchanged. Consequently, if both sides of equation (5) are divided by E_I we have:

$$\frac{E_o}{E_I} = \frac{A}{1 - A\beta} \qquad (6A)$$

where $\dfrac{E_o}{E_I}$ is equal to the total gain of the feedback system, thus:

$$\textit{Total gain of feedback system} = \frac{A}{1 - A\beta} \qquad (6B)$$

Equation (6B) is the fundamental relation for the total gain of any amplifier feedback system. If the type of feedback in question is regenerative, then the value of β used is posi-

tive and equation (6B) remains as it is shown above. If the feedback, however, is degenerative, then the value of β used is negative and the denominator of equation (6B) becomes $1 + A\beta$. In the circuit to be discussed the amount of regenerative feedback voltage is not large and the denominator of equation (6B) will always be less than one (1) for this case whether positive or negative in value.
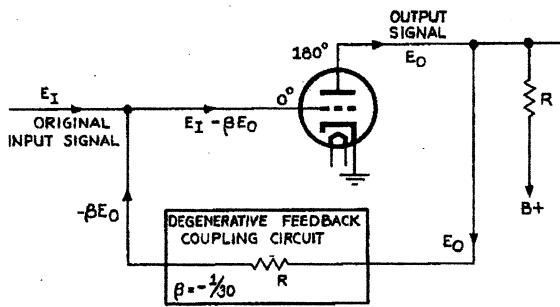

Fig. 5.—In a degenerative feedback circuit, the output signal is 180° out of phase with the input and the fraction β is considered as negative.

As a simple example of how the equation can be used to calculate the gain of a feedback system consider the following:

If the gain A of an amplifier without feedback is equal to 20, (i.e. the output voltage E_o is equal to 20 times the input voltage E_I) and the amount of voltage fed back to the input signal grid is equal to 1/30 of the output voltage, then $A = 20$ and $\beta = \dfrac{1}{30}$. If the feedback is degenerative, as represented by Fig. 5 (the output signal is 180° out of phase with the input signal on the grid), then β will equal $-1/30$ and the gain of the feedback system will be as follows:

$$\text{Total Gain of Degenerative Feedback System} = \frac{A}{1 + A\beta} =$$

(since β is negative)

$$= \frac{20}{1 + 20 \times 1/30} = \frac{20}{1 + \dfrac{2}{3}} = \frac{20}{5/3}$$

$$\text{Total Gain of Degenerative Feedback System} = 12$$

Consequently the gain of the degenerative feedback system is seen to be reduced to 12. In the circuit of Fig. 5, the feedback is degenerative

because the signal existing in the output of the tube is 180° out of phase with that existing at the input to where the signal is fed back.

In the circuit of Fig. 6, two amplifiers are employed to illustrate regenerative feedback. Assuming an input signal of 0° existing on the grid of tube V_1, then the signals existing on the other elements of the tubes will be as shown in Fig. 6. Consequently, the signal fed back from the output tube V_2 to the input of the first stage, tube V_1, will be regenerative. Therefore if β is equal to 1/40 (and positive in nature) and with the gain A of the complete two-stage amplifier system without feedback equal to 20, then:

$$\text{Total gain of regenerative feedback system} = \frac{A}{1 - A\beta},$$

$$= \frac{20}{1 - 20 \times 1/40} = \frac{20}{1 - 1/2} = \frac{20}{1/2}$$

$$\text{Total gain of regenerative feedback system} = 40$$

Consequently the gain of the regenerative system under discussion is equal to 40.
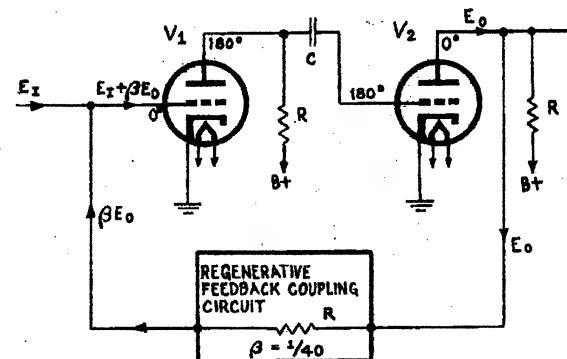
Fig. 6.—In a regenerative feedback circuit, the output signal is in phase with the input and the fraction β is considered to be positive.

All of the feedback circuits to be discussed in this section will be relative to amplifiers, but since oscillators are essentially feedback circuits a few words about them will not be amiss.

In the special section on "Oscillators" it was shown how an oscillator tube returned some of its output voltage to its input circuit, in the proper phase relation, to sustain oscillations. Since this feedback voltage had to sustain oscillations because of the loss in power due to the resistance in the circuit, it had to aid the signal existing at the input to the oscillator and there-

fore had to be regenerative. In this respect, there had to be a proper phase reversal of the signal in the output circuit (i.e. the signal had to undergo a 180° phase shift) in order that this signal be fed back to the input circuit in phase for regenerative feedback. This was accomplished by the transformer network of the tickler-coil oscillator and by the effective auto-transformer networks of both the Hartley and Colpitts oscillators.

Hoffman Model A 301

In the audio system of the Hoffman Models A301, A500, and A501, a 6SQ7 tube is used as a second detector—avc—first audio, and a 6K6GT/G tube as an audio power-output tube. With this system, it was found that there was an undesired amount of audio distortion in the loudspeaker output. Consequently, an inverse feedback arrangment was introduced into the audio system to reduce the unwanted audio distortion. This circuit is illustrated in Fig. 7. The feedback path is from the secondary of the output transformer through the 330-ohm feedback coupling resistor R13, to the cathode of the 6SQ7 tube. The reason why it is negative feedback may not, at first, be readily evident. A few things about this circuit should be understood before the feedback is shown to be degenerative.

By now it should be known that there is a 180° phase shift between the signal existing on the plate with respect to that signal existing on the signal grid. In other words, the signal on the plate is in opposite phase with that existing on the signal grid. Therefore if a signal of zero degrees is assumed existing on the grid of the 6SQ7 tube, then a signal 180° out of phase will exist on the plate of the same tube. The 180° output signal from the plate of the 6SQ7 tube is injected into the signal grid

circuit of the 6K6GT/G tube. Consequently a zero-degree signal now exists on the plate of this latter tube.

Next, the output transformer T5 is wound in such a manner that the signal voltages appearing at points A and D are in the same phase.

Finally, since a zero-degree signal exists on the plate of the 6K6GT/G tube, there also is a zero-degree signal at point A on the primary of T5. In view of these above facts, there is a 0° signal at point D of the secondary of T5.

It is from this point D, that the feedback voltage is fed back to the 6SQ7 cathode. Since a 0° signal exists at point D, it causes the cathode voltage of the 6SQ7 to rise and fall
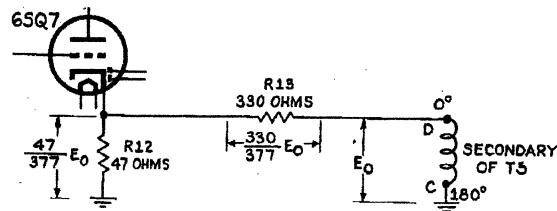
Fig. 8.—Simplified schematic of Fig. 7 illustrating the use of a voltage-divider network, resistors R12 and R13.

in step with the grid signal voltage. As was explained above, this is what happens when a large unbypassed cathode resistor is used; consequently, the signal fed back is degenerative. The 330-ohm resistor R13 limits the amount of feedback voltage returning to the 6SQ7 tube and resistor R13 in conjunction with the 47-ohm cathode resistor R12 forms a voltage-divider network determining the amount of voltage fed back. That is to say resistor R12, in conjunction with R13, determines the value of β used in the previous equations. This is more easily seen in the feedback circuit that is redrawn shown in Fig. 8.

The output signal voltage appearing across
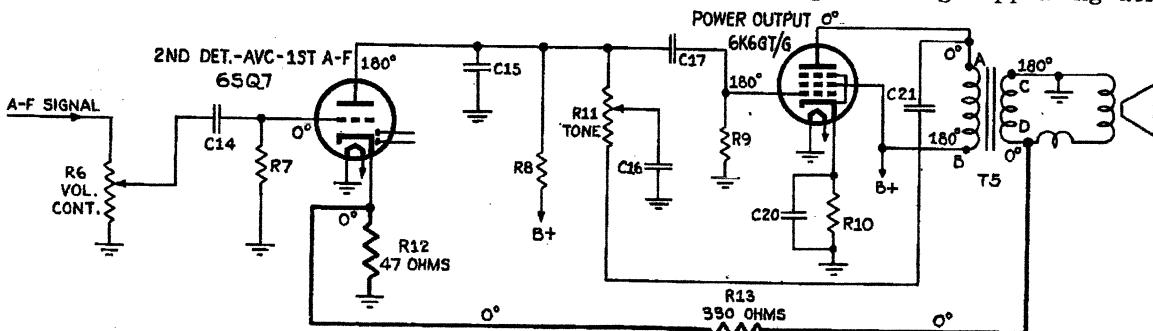
Fig. 7.—Degenerative feedback circuit employed in the audio amplifier of the Hoffman Model A301.

the secondary of the output transformer is designated as E_o. This voltage is applied across the voltage divider network consisting of $R12$ and $R13$; consequently, a certain amount of voltage drop occurs across both $R12$ and $R13$. The voltage drop across $R13$ is given as follows:

$$Voltage\ drop\ across\ R13 = \frac{R13}{R12 + R13} \times E_o$$

$$= \frac{330}{47 + 330} \times E_o.$$

$$= \frac{330}{377} \times E_o$$

$$= .876\,E_o$$

The voltage drop across $R12$ is then given by

$$Voltage\ drop\ across\ R12 = \frac{R12}{R12 + R13} \times E_o$$

$$= \frac{47}{47 + 330} \times E_o$$

$$= \frac{47}{377} \times E_o$$

$$= .124\,E_o$$

The voltage drop across $R12$ can also be found, if the drop across $R13$ is known; that is, the voltage drop across $R12$ is equal to the difference between the total applied voltage E_o and that voltage drop across $R13$. In other words:

$$Voltage\ drop\ across\ R12 = E_o - .876\,E_o$$

$$= (1 - .876)\,E_o$$

$$= .124\,E_o$$

which is the same answer as that above.

That signal voltage drop appearing across the cathode resistor $R12$ is the feedback voltage effectively appearing at the grid circuit of the 6SQ7 tube as an out-of-phase signal. From the above analysis, it is found that β, the fraction of the output voltage fed back, is equal to .124 or about $\frac{1}{8}$. Consequently, the amount of degenerative voltage fed back is equal to about one-eighth of the total output signal voltage appearing across the secondary of the output transformer.

A word of importance relative to the output transformer is necessary, especially if repairs are made on this model and the output transformer has to be removed. As seen in Fig. 7, one end of the secondary, point C, is grounded and the phase of the signal existing at each point established. If a new transformer or one that has been repaired, is installed, it is important that the proper end of the secondary be grounded. If it is not, then instead of the feedback signal being negative, the feedback signal will be positive and this is undesired.

Zenith Model 6D014

In the Zenith Models 6D014 and 6D029, a similar feedback arrangement is used as that discussed in the previous models. A portion of the output voltage from the secondary of the output transformer is fed back to the first audio stage, as shown in Fig. 9. The feedback in this instance is regenerative, and the signal is fed back directly to the grid circuit of the 12SQ7GT first audio stage. The regenerative feedback can be easily seen by tracing a signal from the grid of the 12SQ7GT tube straight through to the speaker.
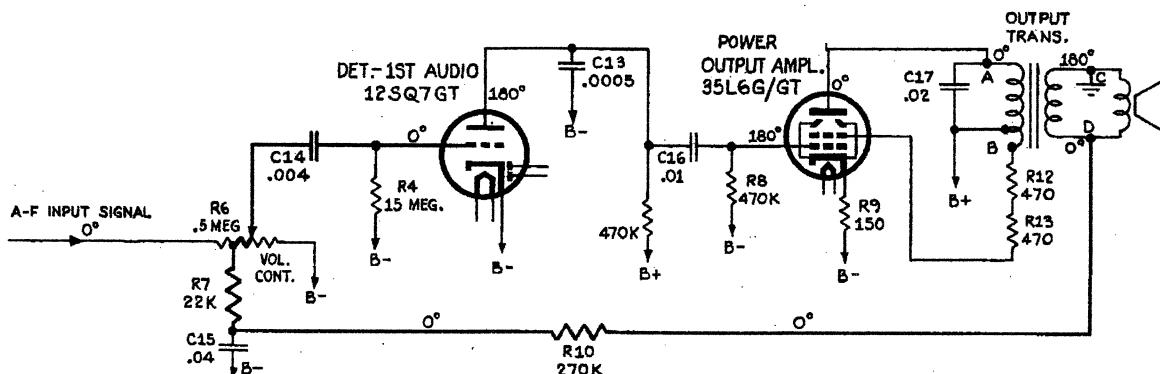

Fig. 9.—In the Zenith Models 6D014 and 6D029 regenerative feedback is used to increase the low-frequency signals.

Assuming that a 0° input signal is injected onto the grid circuit of the 12SQ7GT tube through the volume control R6, then a signal of 180° appears on the plate of the tube due to the phase inversion accomplished within the tube itself. Tracing this signal further, it is found to be coupled to the grid of the 35L6G/GT power output amplifier through the coupling capacitor C16, and most of the signal appears across the grid-leak resistor R8. This R-C coupling arrangement offers negligible phase shift to the 180° signal appearing on the plate of the 12SQ7GT tube and consequently an 180° signal also appears on the control grid of the 35L6G/GT tube. Once again, by virtue of the phase inversion within a tube, a 0° signal appears on the plate of the 35L6G/GT tube. Since point A of the output transformer is common to the plate of the 35L6G/GT tube, a 0° signal also appears at point A.

By virtue of the phase inverting qualities of a transformer, as previously mentioned, a 180° signal exists at point C and a 0° signal at point D of the secondary of the output transformer. From point D a certain amount of voltage is fed back to the grid circuit of the 12SQ7GT tube through the 270,000-ohm feedback limiting resistor R10, the feedback circuit being indicated by the heavy lined path. From resistor R10 the feedback signal currents flow through R7, part of the volume control R6, through C14, and to the grid circuit of the 12SQ7GT tube.

Since the original input to the grid circuit of the 12SQ7GT was a 0° audio signal and the signal fed back to the same grid also of the same phase (zero degree), then it follows that the feedback is regenerative. In other words, the original input and signals fed back are of the same phase and aid each other. This is indicated by the curves of Fig. 2 where curve No. 3 is the resultant input signal, which is seen to be greater than either signal alone.
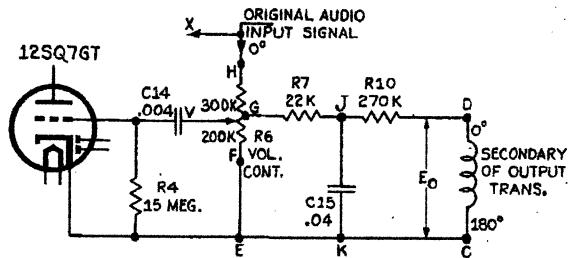
Fig. 10.—Simplified schematic of the circuit of Fig. 9.

The value of the feedback factor β is difficult to estimate from this circuit due to all the components that are relatively in the feedback arrangment. However, if a simplified circuit of the feedback network is redrawn, as shown in Fig. 10, a few points of interest will be noted.

This feedback network is primarily intended to cause regeneration at low audio frequencies; consequently, it is termed a "bass-booster" network. The total output audio signal that appears across the secondary of the output transformer also appears across points D and K of Fig. 10, and the voltage is designated as E_o. This voltage is divided between the voltage divider network of C15 and R10. The reactance of the .04-mf capacitor C15 is so small at the high audio frequencies compared to the resistance of R10 that practically all of the voltage appears across R10. Consequently, a negligible amount of voltage appears across C15 and there is effectively no regeneration at the high audio frequencies. In other words, it can be said that capacitor C15 bypasses the high audio notes to ground because of the small amount of reactance it offers at these frequencies. For instance, at a high audio frequency of 3000 cycles, the reactance of C15 is equal to 1325 ohms. According to this value of reactance only about one half of one percent of the total available output voltage E_o, which is negligibly small, is impressed across C15.

At the low audio frequencies, however, the reactance of C15 increases and a greater amount of the output voltage E_o, is developed across this capacitor. For instance, at 100 cycles the reactance of C15 is eequal about 40,000 ohms. At this frequency the amount of output voltage appearing across C15 is about 14.6% of E_o. This is quite different from the voltage appearing across C15 at 3000 cycles.

Resistance R7 and that section of the volume control between points G and F, are a further voltage divider network to the signal voltage existing across C15, or between points J and E. The amount of resistance between points H and G of the volume control R6 is equal to 300,000 ohms and that between points G and F equal to 200,000 ohms. The 300,000-ohm resistance of the volume control is in series with a high-impedance path, point X, to ground. This total series network is effectively in parallel with the 200,000-ohm section of the volume control. However, the total series net-

work offers too high an impedance, and practically all of the feedback signal current flows through the 200,000-ohm portion of the volume control. Consequently, the effect of the high-impedance path is considered to be negligible. Therefore, the amount of voltage appearing across the 200,000-ohm section of the volume control is equal to about 91% of the total available signal between points J and E. (9% of this latter voltage is dropped across $R7$).

The variable arm V of the control determines how much voltage is to be fed back by tapping off a certain portion of the voltage between points G and F. That voltage appearing across the volume-control arm and point F is impressed across the voltage divider network of $C14$ and $R4$. The reactance of the .004-mf capacitor $C14$ at 100 cycles is about 400,000 ohms, but in conjunction with the 15,000,000-ohm resistor $R4$ very little voltage is dropped across $C14$. In fact, at 100 cycles the amount of voltage drop across $C14$ is about 2½% of the total signal voltage appearing between the variable arm of the volume control and point F. Consequently most of this latter voltage, 97½% of it, appears across $R4$ and is impressed onto the grid circuit of the 12SQ7GT tube.

As a final analysis, it will be interesting to know how much of the output voltage E_o, across the secondary of the output transformer, is impressed across $R4$ at some specific low audio frequency. Since a frequency of 100 cycles was already used and percentage values at this frequency available, therefore, these values will be used again. The amount of voltage appearing across $C15$ at 100 cycles is about 14.6% of E_o, as stated previously. That signal voltage appearing across the 200,000-ohm tap of the volume control $R6$, (between points G and F) is equal to about 91% of that voltage across $C15$. Therefore, .91 $\times$ .146 E_o is equal to .133 E_o, and it can then be stated that 13.3% of E_o appears across points G and F of the volume control. The amount of this latter voltage, 13.3% of E_o, that is available for the voltage divider network of $C14$ and $R4$ is determined by the variable arm (V) of the volume control. In other words, if the variable arm is anywhere between point G and F, then it controls the amount of feedback voltage fed to the voltage divider network of $C14$ and $R4$. If the arm is considered to be at point G, then all of the voltage across points G to F is available

for $C14$ and $R4$. From the previous discussion, $R4$ has a voltage drop that is 97½% of that voltage appearing across points G and F; therefore, .975 $\times$ .133 E_o is equal to about .13 E_o. It can then be stated that the amount of available voltage of E_o to be fed back to the signal-grid circuit of the 12SQ7GT tube, under the above circumstances, is equal to about 13% of E_o. It should be remembered that this voltage E_o is that audio signal appearing across the secondary of the output transformer.

Olympic Model 6-617

In the Olympic Models 6-617 and 6-617U a negative feedback design in the output audio stage is used to increase effectively the bass notes of the set. The circuit for this arrange-
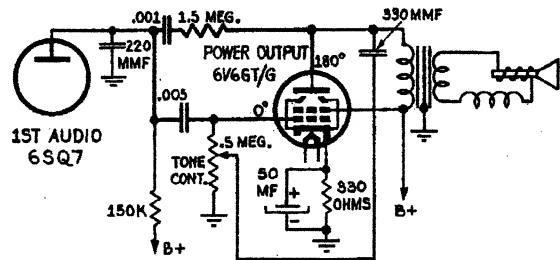
FIG. 11.—Degenerative feedback action is obtained in the Olympic Models 6-617 and 6-617U by feeding the output of the 6V6GT/G back to its control grid over the path indicated by the heavy lines, thus reducing the gain of the high frequencies.

ment is illustrated in Fig. 11. It will be noticed that the feedback is wholly confined within the 6V6GT/G last audio stage. The actual feedback path is from the output of the 6V6GT/G tube (plate) through a 1.5-megohm resistor and .001-mf capacitor feedback coupling arrangement, then through a .005-mf coupling capacitor (that is part of the RC coupling arrangement between the two audio stages) and then onto the 6V6GT/G control grid.

If an assumed 0° audio signal exists on the control grid of the 6V6GT/G tube, then a signal of 180° appears on the plate of the same tube. Consequently, if some of this plate signal is fed back to the grid circuit, the feedback signal will be out of phase with the original signal existing on the grid and the feedback will therefore be degenerative. The .001-mf capacitor, however, offers a varying degree of reactance with change in frequency. As the audio frequency increases, the reactance of the

capacitor decreases and therefore the capacitor offers a more ready path for the high frequencies than for the low frequencies. Since the feedback is degenerative, it follows that the high audio frequencies will be reduced in gain, while the low or bass audio frequencies will remain unchanged. Since degeneration reduces the overall gain of a stage at certain desired high frequencies, then distortion of these high audio frequencies is likewise reduced.

Since the gain at the high frequencies is reduced by the feedback arrangement, the gain at the low audio frequencies is therefore *effectively* increased. In view of this fact a type of feedback arrangement as that shown in Fig. 11 is often termed a "bass-booster" circuit. It will be recalled that the feedback arrangment of Fig. 9 was likewise called a bass-booster circuit, even though that had a regenerative effect, whereas the circuit of Fig. 11 is degenerative. In the former case, the gain at the low frequencies is increased and in the latter case the gain at the high frequencies is decreased; therefore in both cases the gain at the low frequencies is effectively increased.

Sears Roebuck Model 6100

In all of the previous receivers discussed, the feedback circuits analyzed were completely in the audio sections of the sets. In the Sears Roebuck Model 6100, however, a feedback system is used in the i-f section. This feedback circuit is illustrated in Fig. 12. The type of
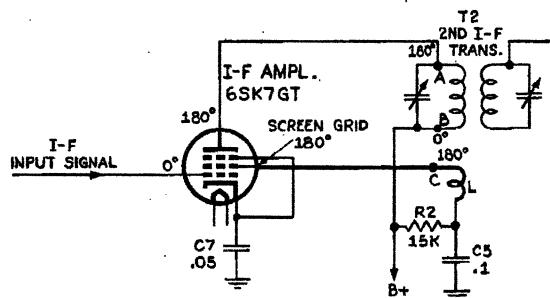
Fig. 12.—The degenerative feedback arrangement in the Sears Roebuck Model 6100 in which the output of the tube is transformer-coupled back to the screen grid.

feedback used is degenerative, although it may not be readily evident. The gain of this i-f section without feedback (the combination of the 6SK7GT tube and i-f transformer) is too high

for the proper operation of this set and consequently a reduction in this gain is desired. The amount of reduction desired is not too great and so a unique method of degenerative feedback is used.

If a 0° signal is assumed to be injected into the control grid of the 6SK7GT amplifier, then a 180° signal appears on the plate of the same tube. Since point A of the second i-f transformer T2, is connected to the plate of the 6SK7GT tube, then a 180° signal also exists at point A. Consequently, a 0° signal exists at point B, the other end of the T2 primary. Below the primary of T2 is another coil, designated as L, which is the feedback element in this circuit. By virtue of the phase inversion between transformer-coupled coils, a 180° signal exists at point C of the feedback coil L. Since point C of this coil is connected to the screen grid of the 6SK7GT tube, then an 180° signal appears on this grid.

Up to this point it is seen that a 0° signal exists on the control grid of the 6SK7GT tube and an 180° signal exists on both the plate and screen grid. It should be remembered, however, that the signal on the screen is obtained by virtue of feedback action. With the feedback in the circuit as shown, the screen grid *effectively* acts as another control grid. In other words, the feedback signal to the screen grid is high enough to make the screen grid act as a control grid and consequently varies the current flowing within the tube. Due to the feedback signal on this latter grid, a signal appears on the plate of the tube which is *opposite in phase* to that signal existing on the screen grid. Since the feedback signal on the screen is at 180°, then a 0° signal will appear on the plate. There already exists, however, an 180° signal on the plate by virtue of the action of the input (control) grid.

The situation is such that due to the 180° feedback signal on the screen grid, there is a 0° signal on the plate which is in opposition to the 180° signal already existing on this plate. Consequently, the feedback signal is degenerative in nature, because it produces a signal on the plate which is opposite in phase to that signal already on the plate, thereby causing a reduction in output voltage, and hence a reduction in gain.

PHASE INVERSION IN PUSH-PULL AMPLIFIERS

To obtain greater and better audio output, as well as to realize several other advantages to be mentioned later, many of the modern radio receivers employ push-pull audio circuits. This art is not new, having been used for many years; however, it has been customary in the past to accomplish the conditions required in such audio systems by the use of transformers. Modern design is often beset by economical limitations, wherein a certain result is desired, yet the financial expenditure must be kept to a minimum. This is done by the use of resistor-capacitor combinations instead of transformers, so that although the number of components used is actually increased, it is done at much less cost than if transformers were used. To permit the application of this arrangement, a circuit combination known as "phase inversion" is employed.

To appreciate properly the significance of phase inversion, a general understanding of push-pull operation must be had; consequently, we shall spend a few minutes in the discussion of this circuit as the preface to the ultimate purpose of this description. In brief, push-pull operation consists of a pair of triode or multi-element vacuum tubes, either individual or contained in a single envelope and so arranged electrically that signal voltages of equal magnitude but of opposite "phase" are applied simultaneously to the individual signal grids. For the moment, the manner in which these signal voltages are obtained in the grid circuit is not important, but we shall assume that they are supplied by a center-tapped transformer as shown later on. Whatever the grid bias applied to the grids, it is the same for both.

The plate circuit of the push-pull stage also connects to a center-tapped transformer, so that like values of operating plate voltage are applied to the plates of the tube. The two halves of this plate transformer primary are identical in electrical characteristics, so that electrical symmetry or balance exists in the plate circuit and likewise in the grid circuit. This is essential in all push-pull circuits in order to realize the advantage of the system. Stated differently, we might say that the first require-

ment in all push-pull circuits is that like grid voltages, operating, as well as signal, exist on the grids of the tubes.

Having established circuits and voltage symmetry, the next consideration is the action which takes place in this amplifying stage. Doubtless you recall one of the paramount fundamental actions which takes place in all triode or multi-element vacuum tubes as amplifiers, this being the so-called "phase inversion" accomplished between the grid and plate circuits. To present this part of the discussion in this limited space properly, it will be well to review an interesting condition associated with amplifying vacuum tubes.

In general, amplifying vacuum tubes are used with either no negative operating grid bias or with some value of negative operating grid bias. Now if a signal voltage is applied momentarily to the control grid, it will change the voltage relationship between the grid and its electron emitter, either making the grid more or less negative depending entirely on the instantaneous polarity of the signal voltage. This action often is referred to as *"increasing"* or *"decreasing"* the grid potential; by *"increasing"* being meant that the grid is made *less* negative, and by *"decreasing"* is meant that the grid is made *more* negative. These are the references to be used in this discussion and in order to make them clear, we felt that this brief fundamental explanation was necessary.

Referring once more to the "phase inversion" which takes place between the grid and plate circuits, this comes about in the following manner: when the grid voltage is increased momentarily, the plate current is increased and the voltage drop across whatever conducting devices are located in the plate circuit likewise increases. Since the plate voltage supply source is constant in its output, the voltage momentarily effective at the plate is decreased. Since the increase in grid voltage represents a "positive" signal voltage, the momentary decrease in plate voltage represents a "negative" signal in the plate circuit. Of course, the reverse is true, that when the grid voltage momentarily decreases, representing a "negative" signal voltage, the plate current decreases and the

voltage drop across whatever conducting element is in the plate circuit, likewise decreases, thus raising the momentary voltage effective at the plate. Therefore, a "negative" signal voltage at the grid produces a "positive" signal in the plate circuit. This action is identified broadly as "phase inversion" between the grid and plate circuits.

Referring once more to the push-pull system, two simultaneous actions are found in the control-grid and plate circuits. In the control-grid circuit, each grid receives a like value of signal voltage; but when one grid is positive, the other grid is negative. Stated differently, the two signal voltages applied to the grids are 180° out-of-phase at every instant but equal in magnitude. In turn, the two signal voltages developed in the plate circuit likewise are 180° out-of-phase with each other and also with their respective grid voltages. Because of the symmetry in the plate circuit, the two out-of-phase signal voltages developed across the respective halves of the transformer primary are equal in value.

In order to accomplish these characteristic features of amplification most economically, phase inversion is employed ahead of the output stage. In Fig. 1 is shown a simplified schematic common to practically all push-pull circuits. In this circuit, the phase inversion is seen to be already accomplished, and that the voltages applied to the two grids are equal in strength and opposite in polarity or 180° out-of-phase (note the + and — signs on the control grids). This means that at the time a

maximum positive audio signal is on the control grid of tube V_1, a maximum negative audio signal is on the control grid of tube V_2. In order to make the rest of the system function properly, the two tubes used have to be identical in electrical characteristics. The same supply voltages must be impressed on each screen grid (if the tube is a tetrode or a pentode) and also both cathodes have to be above ground (if self-biasing is used), by the same amount.

Phase Inversion Action

The phase-inverted signals are controlled in such a manner that the instant the signal voltage on the control grid of tube V_1 is a *positive maximum*, the signal voltage on the control grid of tube V_2 is a *negative maximum*. In this respect the a-c plate current of tube V_1 will be increasing to its maximum and at the same time the a-c plate current of tube V_2 will be decreasing to its lowest value. These a-c plate currents have certain relations toward each other that are beneficial to the set. First of all, these currents are flowing in such a manner that they will produce voltages in the *secondary* of the output transformer that are in the *same* direction, thus *increased* power output is obtained. Secondly, the wave shapes of these output voltages are such that when they are added together no even harmonics will be in the combined output voltage and hence there is elimination of the second and all other even order harmonic *distortion*. The reason this is possible is because the sum of the individual output voltages will have shapes that contain only positive and negative half cycles that differ only in sign, not in amplitude. We find that the d-c output plate currents from the push-pull tubes will be flowing in opposite directions in their particular half of the tapped *primary* winding of the output transformer. Due to this means of output current flow, d-c saturation of the transformer core is prevented. Because the plate currents magnetize the core in opposite directions, the magnetizing effects cancel each other, thereby preventing d-c saturation of the core, thus a greater amount of plate current can be drawn through the push-pull output transformer than could be obtained with a conventional output transformer using a single output tube.

Any a-c hum voltages that find their way into the push-pull plate circuits will be sys-

tematically eliminated and no hum will appear in the output, because the a-c hum currents that flow in both sections of the primary of the output transformer will balance each other. The reason for this is that when the a-c hum voltages are in the plate circuit they are of the *same phase* and instead of becoming additive, they balance each other out by producing currents of *opposite* directions in the primary of the push-pull output transformer. This, of course, applies to a-c hum produced only in this push-pull stage, as mentioned, because if the a-c hum appears in the other stages of the set, it will be incorporated as part of the audio signal, and therefore, will be heard in the output.
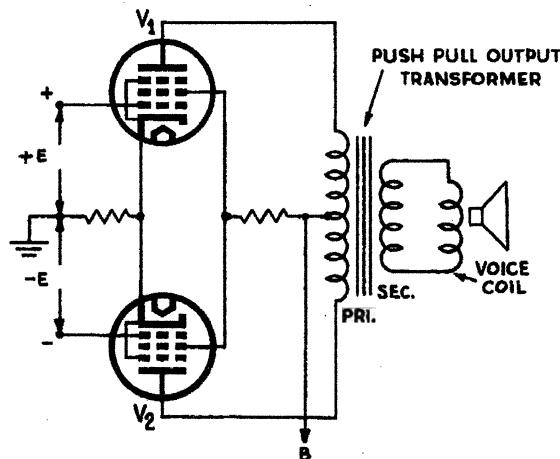
Another beneficial feature, due to the balancing effect in the output transformer circuit, is that there is no *fundamental* audio frequency flowing through the source of plate supply. Therefore, no degeneration can occur and consequently, a cathode bypass capacitor is not needed to shunt the cathode biasing resistor. It may be found, however, that some sets do use a cathode bypass capacitor in their push-pull circuits. The reason it is used at all, is to make doubly sure that if there be any discrepancy whatsoever in the matching of component parts for push-pull circuits, the capacitor would bypass any undesired fundamental signal currents that might possibly flow through the cathode biasing resistor. Therefore, when the capacitor is used, it is employed as a precautionary means only.

This type of audio power-output amplification (push-pull) is used in a very great many radio sets because of the following advantages:

a) Reduction of harmonic distortion by cancelling out the second harmonic and all even order harmonics.

b) Greater power output for the same voltage input than that obtainable with a single tube.

c) Larger plate currents can be drawn through the push-pull output transformer than ordinarily can be obtained in a single tube output stage.

d) Reduction of a-c hum voltage that appears in the push-pull stage only.

e) Filtering by a cathode bypass capacitor across the cathode biasing resistor is not needed to avoid feedback difficulties as usually is employed in the class A push-pull output amplifier.

In order to make as near an ideal operating push-pull system as possible, it is most important to obtain good phase reversal to the input of the system. In the past, the most common method of obtaining phase reversal was to have an *input* push-pull transformer between the output of the last a-f stage and the input of the push-pull system, as shown in Fig. 2. In this way, the audio signal from the voltage amplifier is impressed across the primary of the input push-pull transformer and coupled across the center-tapped secondary. Since the secondary is center-tapped, a phase reversal occurs in the signal appearing across the secondary such that the input to one of the control grids of the push-pull tubes is 180° out-of-phase with the other. In modern receivers, however, the tendency is to omit the input push-pull transformer, mainly because of the great amount of space it occupies; therefore, numerous circuits in push-pull operated receivers accomplish their phase reversal by some other means. In most of these circuits some form of *phase-inversion* tube is used, and this tube is commonly known as a *phase inverter*. In the following paragraphs numerous phase-inverter circuits will be discussed that are encountered in many of the radio sets on the market today.

In order to understand phase-inverter circuits, the operation of a vacuum tube with respect to the relationship between the voltages applied to the grid and current in the plate circuit, must be understood. Because of tube characteristics, a signal voltage applied to the grid of a tube will cause a current flow in the plate circuit of that tube, which will at all times be 180° out-of-phase with the signal voltage.
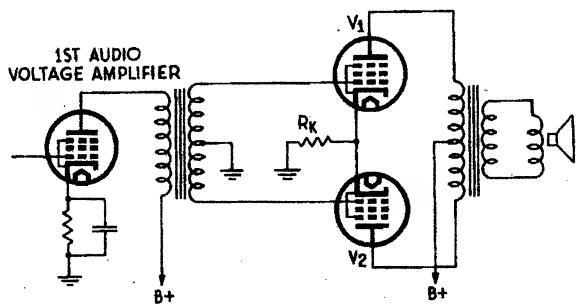

Fig. 2.—The common push-pull circuit employs an input transformer with tapped secondary as well as an output transformer, as in Fig. 1.

In the accompanying discussion of phase inversion circuits, it will be easier at times to refer to signals as being plus (+) and minus (—) with respect to phase reversals and at other times to refer to these signals as having a definite phase reversal of 180°. In this latter aspect, the phase reversal referred to will be between 0° and 180° for the sake of simplicity.

Motorola Model CR6

In the Galvin Motorola Model CR6, a 6SQ7GT tube is used as a phase inverter, as shown in Fig. 3. The inverter tube, which is a duo-diode triode has both diode plates grounded and only the triode portion is used. Since the idea for proper push-pull operation is to obtain equal but out-of-phase voltages applied to the control grids of the 6V6GT push-pull power amplifiers from the preceding audio stages, this receiver employs the following circuit:

From the plate of the first-audio tube (the triode portion of the upper 6SQ7GT tube in Fig. 3) a signal voltage is impresssed onto the control grid of the top 6V6 power amplifier through C23. Assuming this voltage to be zero degrees, then it can be said an audio signal of zero-degree phase relation is impressed across the grid-leak resistance network as used by the top 6V6 tube. Now since the grid-leak resistors R20 and R21 of the 6V6 power amplifiers are not equal, then some of this zero-degree signal (from the top 6V6 tube) effectively is tapped off this grid-leak resistor combination and ap-

plied as an input signal to the grid of the 6SQ7GT phase-inverter tube. By virtue of the tubes properties, the input zero-degree signal is amplified somewhat and it appears on the plate of the phase-inverter tube as an amplified signal, but out-of-phase by 180° with the input to its grid. This 180° signal is in turn impressed across the control grid of the lower 6V6 push-pull tube through capacitor C25. Therefore it is seen how the input to the control grids of the 6V6 push-pull tubes are out-of-phase with each other. The factors determining the equality of the magnitude of the a-c voltages applied to the grid of the 6V6's are the values of R20, R21, and the amplification of the phase-inverter tube. The grid of the phase-inverter tube receives a certain amount of audio signal from the grid circuit of the top 6V6 tube according to the ratio of R20 to R21. In this case, the difference between R20 (220,000 ohms) and R21 (270,000 ohms) being 50,000, this is the determining factor of the amount of voltage to be impressed across the grid of the phase-inverter tube. The amount of this input voltage together with the gain of the inverter stage, makes the input signal voltage to the lower 6V6 push-pull tube equal to that input signal voltage of the top tube.

Spiegel Model W-106

In the Spiegel Model W-106, Fig. 4, the phase inversion is similar in most respects to
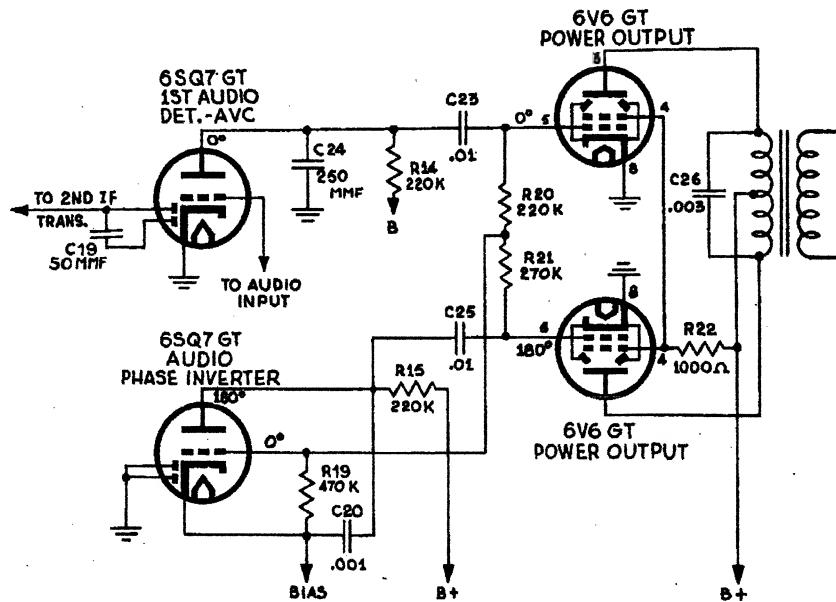
Fig. 3.—The triode portion of the lower 6SQ7GT tube is used as a phase inverter in the Motorola Model CR6. A portion of the input to the upper 6V6GT tube is fed to the grid of the phase-inverter tube. The output signal of this tube is 180° out of phase with the input to the upper 6V6GT tube, but the voltages of these two signals are equal. Hence, the input signals to both power output tubes are the same amplitude but opposite in phase.

the Motorola set just discussed. In this Spiegel set, the phase inversion is accomplished similarly by an inverter tube, type 76.
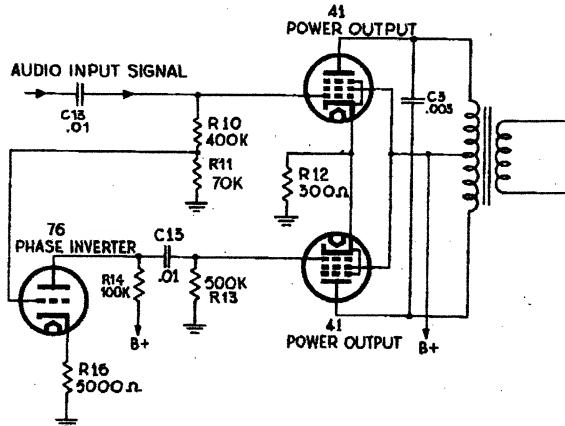

FIG. 4.—The output of the 76 phase-inverter triode is impressed on the control grid of the lower 41 power-output amplifier tube in the Spiegel Model W106 receiver.

The grid circuit of the 76 phase-inverter tube is tapped across a certain portion of the grid-leak resistance (*R10* plus *R11*) of the top 41 push-pull tube. The grid of the phase-inverter tube, therefore, receives a certain amount of the signal impressed across the grid of the top push-pull amplifier. This signal is in turn amplified and undergoes a phase reversal between the·grid and plate of the phase-inverter tube. Therefore, the output signal from the plate of the phase-inverter tube is 180° out-of-phase with that signal at the input of the top 41 push-pull tube and this signal reversed in phase is now impressed across the control grid of the lower 41 push-pull tube.

The controlling factors that determine the equality of the applied out-of-phase voltages to the push-pull grids are the ratio of *R10* to *R11* and the amplification of the 76 phase-inverter tube. That is to say, the input to the 76 tube multiplied by the gain of this inverter stage should be approximately equal in voltage to that input voltage to the top 41 push-pull tube.

Zenith Model 12H090

In the push-pull circuit of the Zenith Model 12H090, as shown in Fig. 5, the inverter system is slightly different from those previously discussed. In this system, the audio output from the plate of the first-audio tube is impressed onto the grid circuits of the 6J5GT phase-in-

verter tube and the lower 6V6GT push-pull tube No. 2.

The amounts of audio signal impressed across the grid circuits of these tubes are different, that amount going to the phase-inverter tube being controlled by the voltage-divider network *R29*, *R27*, and *R32*. In other words, the total audio-output voltage from the first-audio stage is impressed across *R29*, *R27* and *R32* to ground. That drop occurring across *R27* (47,000 ohms) in series with *R32* (33 ohms) is impressed onto the grid of the phase-inverter tube. This drop is

$$\frac{47,000 + 33}{330,000 + 47,000 + 33}$$

or approximately one-eighth of the total available audio voltage. Assuming the phase relation on the grids of both the phase-inverter tube and No. 2 6V6 push-pull tube to be a zero-degree signal, then that amplified signal appearing at the plate of the phase-inverter tube is 180° out-of-phase with the signal on the phase-inverter grid; therefore it is also 180° out-of-phase with the signal on the grid of No. 2 push-pull tube. This amplified signal from the phase-inverter plate is impressed across the control-grid circuit of the top No. 1 6V6 push-pull tube. Consequently, the push-pull control grids of the 6V6 tubes have 180° out-of-phase signals impressed across them.

This set was designed so that a certain amount of input signal is fed into the phase-inverter tube (about one-eighth of that impressed across the No. 2 push-pull tube) and this signal would be amplified by the tube. The input signal in conjunction with the amplification of the phase-inverter tube made the signal voltage on the grid of the No. 1 6V6 push-pull
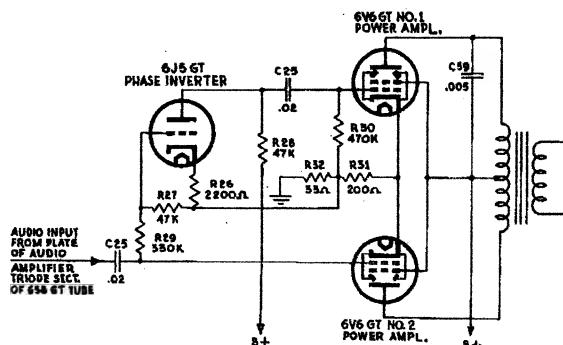

FIG. 5.—The output of the 6J5GT phase inverter is applied to the control grid of the upper 6V6GT power amplifier in the Zenith Model 12H090.

tube approximately equal in *magnitude* to that impressed across the grid of the No. 2 6V6 push-pull tube but 180° out-of-phase with it. In all of the sets using a phase-inverter tube it must be remembered that the amplification of the tube is controlled by the voltages applied to all of its terminals and, therefore, the voltages have to be adjusted for the desired amplification needed.

RCA Model Q34

In the RCA Model Q34 as shown in Fig. 6, a dual-purpose tube is employed which functions as a phase inverter and one section of the push-pull system. This tube is the 6AD7G. The pentode section of this tube is used as one part of the push-pull system and it is similar in electrical characteristics to a 6F6G tube. Since the push-pull output tubes have to be as identical in electrical characteristics as possible, this is the reason why the second half of the push-pull system uses a 6F6G tube. The triode section of the 6AD7G tube functions as a phase inverter. The complete system functions as follows:

Assuming a positive input signal is coming from the plate of the first-audio tube, it is then impressed across the resistors $R25$ and $R26$. This in turn impresses the total positive signal on the control grid of the pentode section of the 6AD7G tube. The grid of the triode phase-inverter section of the 6AD7G tube is tapped between the two grid resistors $R25$ and $R26$. The inverter grid then receives a certain percentage of the positive signal applied across $R25$ and $R26$. The amount impressed is determined by the ratio of $\dfrac{R26}{R26 + R25}$ times the
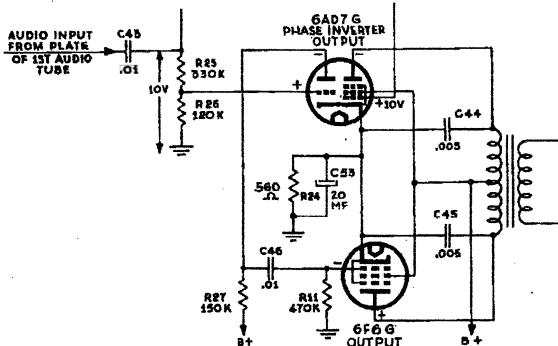
Fig. 6.—In this RCA Model Q34, the triode portion of the 6AD7G output tube is used as a phase inverter.

available signal voltage — and it is measured as the voltage drop across $R26$. The positive signal on the phase-inverter grid undergoes a phase reversal within the tube and there appears an amplified (negative) signal on its plate. This negative signal is then impressed across the grid-resistor $R11$ of the 6F6G tube through $C46$. In this manner, the control grid of the 6F6G tube receives a negative signal as compared to the positive signal on the control grid of the 6AD7G, the other half of the push-pull network. The application of equal, but out-of-phase, voltages to these push-pull control grids are derived similar to the previously discussed push-pull inverter circuits. This can be better illustrated, however, with a few simple figures.

Assuming the signal voltage applied across the grid-leak resistors $R25$ and $R26$ to be 10 volts, then 10 volts (positive) audio signal appears on the control grid of the pentode push-pull section of the 6AD7G. The voltage appearing on the grid of the phase-inverter section of the 6AD7G tube is, by virtue of the voltage drop across $R26$:

$$\frac{R26}{R25 + R26} \times 10 = \frac{120K}{120K + 330K} \times 10 = \frac{120}{450} \times 10 = 2\tfrac{2}{3} \text{ Volts}$$

In order for there to be a 10-volt (negative) signal on the control grid of the 6F6G tube, then the amplification of the phase-inverter triode section of the 6AD7G tube must be:

Amplification $\times$ $2\tfrac{2}{3}$ = 10 Volts or

$$\text{Amplification} = \frac{10}{2\tfrac{2}{3}} = 3\tfrac{3}{4}$$

Therefore, with a grid-leak combination of $R25$ and $R26$ as shown in Fig. 6 and with an assumed 10 volts on the grid of the 6AD7G pentode section, the potentials of the triode inverter section is such that the tube will have a gain of $3\tfrac{3}{4}$.

Brunswick Model BJ6836

In the Brunswick Model BJ6836, a duo-triode 6SN7GT tube serves as a second-audio voltage amplifier and a phase inverter, as shown in Fig. 7. The second-audio voltage amplifier is resistance coupled from the first-audio voltage amplifier (a 6J5 tube). Both triode units

of the 6SN7GT tube have practically the same electrical characteristics.

From the plate of the audio-amplifier section of the 6SN7GT, the audio signal is impressed across the grid of the upper 6V6GT push-pull amplifier. Part of that audio signal that is impressed across the total grid-leak resistance, namely that signal voltage across the 100,000-
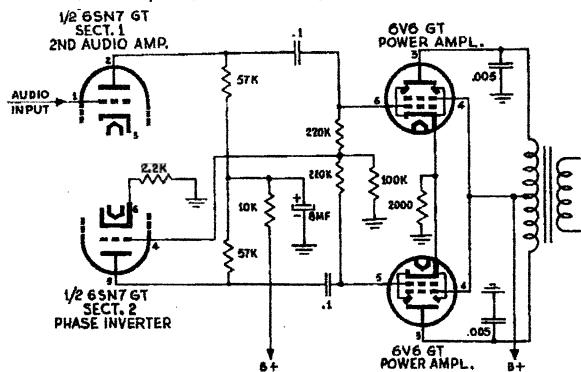

FIG. 7.—In the Brunswick Model BJ6836 one-half of the 6SN7GT tube functions as the phase inverter, the other half being the second a-f amplifier.

ohm part of the grid-leak resistance, is impressed onto the grid of the inverter section 2 of the 6SN7GT tube. The signal appearing on the plate (pin 5) of the inverter section of the 6SN7GT is opposite in phase of that signal on its grid. This phase-inverted signal appearing on the plate of the inverter section is then impressed across the grid of the lower 6V6GT push-pull amplifier tube.

Therefore, it is seen easily how the phase inversion in such a system is accomplished. The application of equal, but out-of-phase voltages on the control grids of the push-pull tubes is controlled, as usual, by the amount of signal-voltage drop across the 100,000-ohm grid-leak resistor and the amplification of the inverter section of the 6SN7GT tube. This voltage (which is impressed across the grid of the inverter) multiplied by the amplification of the inverter section of the 6SN7GT tube is that signal voltage impressed across the lower 6V6GT push-pull tube. This amplified signal has to be equal in voltage to that output from the audio-voltage amplifier section of the 6SN7GT tube, which in turn is impressed across the top 6V6GT push-pull tube. The accomplishment of equal voltages is more easily obtained in this instance since both halves of the 6SN7GT tube have equal plate voltages applied to them, making the output characteris-

tics of each half of the 6SN7GT tube dependent upon the signal input to their separate grids.

General Electric Models X181 and XC181

In the General Electric Models X181 and XC181 series there is a unique phase-inversion circuit. The method of obtaining equal and out-of-phase signal voltages on the control grids of the 6V6 push-pull tubes will be understood easily if the instantaneous polarities of the signal voltage are considered. First it will be best to understand how the out-of-phase signal voltages reach the control grids of the 6V6 push-pull tubes, and then, the method of making these out-of-phase voltages equal will be considered.

Referring to Fig. 8, the instantaneous signal current is shown by the arrows for a specific instant of time, flowing from the plate of the 6J5 phase-inverter tube. The signal current flows from the plate through the plate load resistor $R22$ and back to the cathode, flowing through the $R21$ and $R20$ cathode resistors to complete its path. Current flowing toward a resistor, $R21$ for example, makes the grounded end negative, and after flowing through the resistor, makes the upper end positive with respect to the grounded end. Therefore, it can be seen that with the signal current flowing, as shown, a negative potential is on the plate side of the loading resistor of the 6J5 tube and a pos-
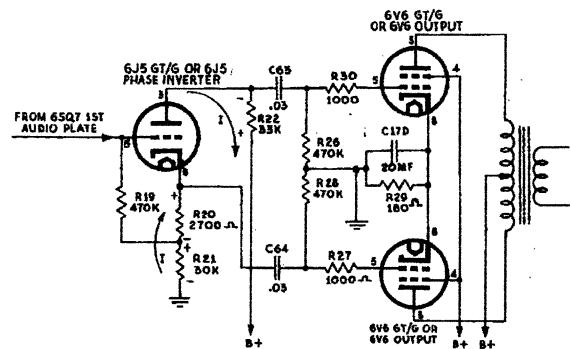

FIG. 8.—The signal impressed on the grid of the upper 6V6 tube through $C63$ and $R30$ is 180 degrees out-of-phase with the signal applied to the grid of the lower 6V6 tube through $C64$ from the cathode resistors $R20$ and $R21$, in this General Electric Model X181, XC181.

itive potential on the other side. Similarly, the potentials on the cathode resistors $R20$ and $R21$ are designated as such due to the returning

signal current. Consequently, there exists a *negative* signal across the load resistor *R22* of the 6J5 phase-inverter tube and at the same time a *positive* signal exists across the cathode resistors. In other words, out-of-phase signals are across the plate load and the cathode load *at any given instant of time.* Therefore, from the plate of the 6J5 phase-inverter tube, the signal is impressed across the control grid of the top 6V6 push-pull tube through *C63*, and from the cathode of the 6J5 phase-inverter tube, the signal (appearing across the cathode resistors) is impressed across the control-grid circuit of the lower 6V6 push-pull tube. In this way, out-of-phase signals are applied to the control grids of the 6V6 push-pull power-output tubes.

Since the signal current flows through both the plate-load resistor *R22* and the cathode resistors *R20* and *R21* (there is no bypass capacitor across the cathode resistors), it becomes feasible to obtain equal signal voltages on the control grids of the push-pull 6V6's. If the plate-loading resistor and the total cathode-loading resistance are made equal then the signal current flowing through both loads will cause a voltage drop in each load that will be equal to each other. The respective out-of-phase, but now equal, voltages are impressed across the control grids of their respective push-pull tubes.

Two Types of Coupling Used

A few interesting things are worth noting about this 6J5 phase-inverter tube. First of all, it uses two types of coupling, namely, the regular resistance-capacitance coupling (*R22, C63*) in the plate circuit of the 6J5 tube and *cathode coupling* (utilizing *R20, R21,* and *C64*) in the cathode circuit. This latter type of coupling is relatively new in the field of broadcast receivers and finds one of its greatest uses in the phase-inverter circuit as described here. Due to this latter method of coupling, the gain of the 6J5 stage can never be greater than unity (one) and at most it can only *approach* such a value because a gain of one, for such a phase-inverter stage (sometimes called a "phase-splitting network"), only is realized ideally. This can best be explained by noting a few facts about the effects of the cathode resistor.

First, when any bypass capacitor is removed from across a cathode-biasing resistor, a signal voltage appears across the cathode-biasing resistor, which is impressed effectively onto the grid circuit of the same tube. This signal is out-of-phase with the signal already existing on the grid thus causing degeneration. In other words, we have inverse feedback. (See section on "Feedback".) The actual *effective* signal voltage now appearing on the grid of the tube is the difference between the original input signal and the signal appearing across the cathode-biasing resistor. In the circuit of Fig. 8 the same is true; degenerative feedback occurs in the 6J5 phase-inverter tube. Since the *effective* signal voltage appearing between the grid and cathode of this tube is equal to the input signal voltage *less* the a-c voltage drop across the cathode resistance, then this voltage drop (which is one part of the output voltage) cannot be greater than the original input voltage. Since the output of the 6J5 tube (that across *R20* + *R21* and *R22* since they are both the same) is less than the input to the same tube, then the gain can never be greater than unity. Even though the gain is less than one, degenerative feedback benefits such a circuit. Since the grid of the 6J5 tube is returned to the junction of the two cathode resistors *R20* and *R21*, through its grid-leak resistor *R19*, then the bias on the tube is due only to the d-c voltage drop in resistor *R20*.

As was stated before, for equal output voltages, the plate-load resistance and the total cathode-loading resistance should be equal. If Fig. 8 is examined, it appears as though this were not so because the plate-load resistor *R22* is 33,000 ohms and the total cathode resistance is 32,700 ohms, (*R20* + *R21*). There appears to be a discrepancy of 300 ohms but if one realizes that the values of resistances are standard values, it will be difficult to get an exact duplicate of two resistances *R20* and *R21* adding up to 33,000 ohms. The closest values obtainable for matching load were used. However, with the tolerance of 10% that is common to most resistances used today, it can be considered a close match for equality of loads because the 300-ohm discrepancy constitutes only 1% of the total 30,000 ohms load as compared to the 10% tolerance allowed.

Sparton Model 7-46

In the Sparks-Withington, Sparton Model

7-46, etc., as shown in Fig. 9, there a type of phase-inversion circuit is incorporated that does not use a separate phase-inverter tube. The phase-inversion and the means of obtaining equal voltages is accomplished by the 6F6G/GT push-pull tubes themselves.
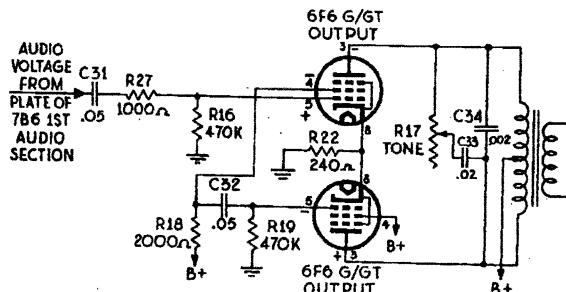

FIG. 9.—The screen-grid current of the upper 6F6G/GT push-pull amplifier develops a voltage drop in *R18*, which voltage is 180 degrees out-of-phase with the input to the upper tube from *R27*; thus, the signal input to the lower 6F6 tube control grid has been reversed 180 degrees.

Upon examination of this circuit, it will be noted that audio voltage from the plate of the first-audio section of a 7B6 tube is impressed onto the grid circuit of the top push-pull 6F6G/GT tube. With a constant supply voltage on both the screen grid and plate of the top 6F6 tube and with a varying input signal, both the plate and screen-grid currents of this tube increase and decrease at the same time with respect to the input signal. Accordingly, assuming a positive input signal on the control grid of the top 6F6 tube, a 180° phase reversal will occur within the tube itself and a negative signal will appear on both the plate and screen of the tube. From the screen grid of this top 6F6 tube part of the negative signal voltage is injected onto the control grid of the lower 6F6 push-pull tube. In this manner, out-of-phase signal voltages are impressed across the control grids of the push-pull amplifier tubes.

In order to have the signal voltages that are impressed upon the push-pull control grids equal in magnitude, then the signal-voltage output from the screen grid of the top 6F6 tube (that signal voltage impressed across the lower 6F6 tube) must equal that signal voltage on the grid of the top 6F6 tube. This is accomplished by having the proper loading resistor on the screen grid of the top 6F6 tube. In the circuit under question, the screen loading resistor *R18* is 2000 ohms and the a-c signal voltage drop across this resistor is just enough to equal that

signal voltage on the control grid of the top 6F6 tube. The a-c signal drop across the 2000-ohm resistor is then impressed onto the control grid of the lower 6F6 tube.

Minerva Model W-117-3

In the Minerva Model W-117-3, the phase-inversion is obtained within the push-pull system itself, similar to the way it was accomplished in the preceding set. As seen in Fig. 10, the Minerva set uses two 50A5 tubes as their push-pull power-output section of the set. The audio-signal input to the grid of the top 50A5 tube is received conventionally from the plate of the first-audio section of a 6SQ7 tube.

Assuming a positive input signal on the control grid of the top 50A5 tube, then a negative signal appears on the plate of the same tube by virtue of the phase-inverting qualities within the tube itself. Some of this negative signal appearing on the plate of the top 50A5 push-pull tube is taken off the plate and impressed onto the control grid of the lower 50A5 push-pull tube. The path this signal takes is from the plate of the top 50A5 tube through the 470,000-ohm resistor and then across the R-C combination of a 100,000-ohm resistor in
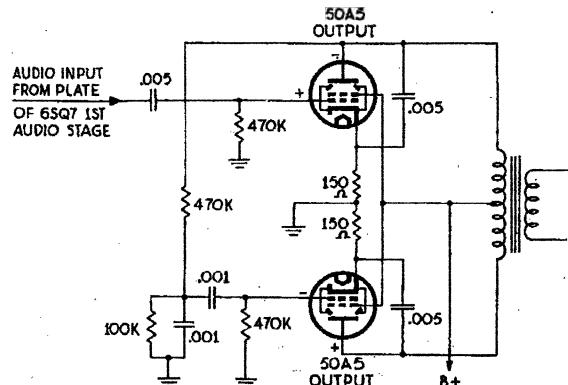

FIG. 10.—A portion of the plate current from the upper 50A5 output tube in the Minerva Model W117-3 is fed through the R-C combination, (100,000 ohms and .001 mf) to ground and thence to the control grid of the lower output tube, thus reversing the input through 180 degrees.

parallel with the .001-mf capacitor. The negative signal is then impressed across the control grid of the lower 50A5 tube to give an out-of-phase signal as compared to that on the control grid of the upper 50A5 push-pull tube.

The amount of signal voltage applied to the control grid of the lower 50A5 tube is determined by the 100,000-ohm resistor, the

470,000-ohm resistor and both .001-mf capacitors. The signal voltage developed across the parallel combination of the 100,000-ohm resistor and .001-mf capacitor is in turn developed across the R-C coupling combination of the .001-mf coupling capacitor and 470,000-ohm grid-leak resistor.

An illustration will make this somewhat clearer. In Fig. 11, the input circuit to the lower 50A5 push-pull circuit is redrawn. Assuming that an instantaneous signal voltage e is developed across points A to B, then this same voltage also exists across points A to D since points B and D are both at ground potential. Between points A and D two components are in series, namely C_b and R_c. From the junction between C_b and R_c a lead goes to the control grid of the lower 50A5 tube. Therefore, C_b and R_c represent a voltage-divider network for the total signal voltage appearing across them both and that part of this instantaneous signal voltage e that is impressed across the grid of the tube, is the drop across the resistor R_c. In other words, the capacitor C_b is so chosen that its reactance, compared to the resistance of R_c, will be such that the voltage developed across the 470,000-ohm resistor R_c is equal to the signal voltage applied to the control grid of the top 50A5 tube.

The reason why emphasis was made previously upon the R_c-C_b combination is that the capacitor C_b is *small* (.001 mf) *compared* to those used in other sets and, therefore, its reactance is very much higher, causing a large enough drop in the signal voltage applied, to be considered as part of a voltage-divider network. However, since the signal voltage represents a varying degree of different audio frequencies and since the reactance of the

capacitor C_b will vary with frequency, the voltage drop caused by this reactance will vary also. In this respect, the audio-response curve will not have as level a characteristic as possible, because at low audio frequencies the reactance of C_b is higher than that at high audio frequencies and the curve for the audio-response characteristics will show a definite decline at the low-frequency end. In order to compensate for this unevenness in audio response, a second .001-mf capacitor C_i is used and it is shunted across the 100,000-ohm resistor R_i. The compensation is seen readily when one considers the total impedance offered by the R_i-C_i network at different audio frequencies compared to the reactance of C_b at these same frequencies. At low audio frequencies, the capacitor C_b has a sufficiently high reactance to cause a decided drop in the response curve at the low end, and the insertion of the capacitor C_i compensates for this drop in response. Since the total available audio signal to the control grid of the lower 50A5 push-pull tube is determined by the total impedance of the R_i-C_i parallel network, the available voltage also varies since the impedance of this network changes with varying audio frequencies due to the reactance of the capacitor C_i changing with frequency. As the frequency *goes up* the reactance of C_1 goes down, and the resistor R_i is in parallel with a lower reactance at higher audio frequencies. The parallel combination of R_i-C_i then offers an overall impedance that decreases with increase in frequency, since any two impedances in parallel offer a total impedance smaller than the smallest impedance. Since the total impedance of R_i-C_i decreases with frequency, then the signal voltage drop across the R_i-C_i combination, the greatest drop being determined by the highest amount of impedance offered, also decreases with increase in frequency. Although the tendency of this type of circuit is to reduce the overall gain available in the input to the 50A5 lower push-pull tube, the frequency response becomes flattened. This drop in gain is, however, satisfactory since the values of all components are so chosen that the input to the control grid of the lower 50A5 tube equals that signal voltage on the control grid of the top 50A5 tube. It will be found that when the capacitor C_b (in conjunction with R_c) discriminates against low audio frequencies, the
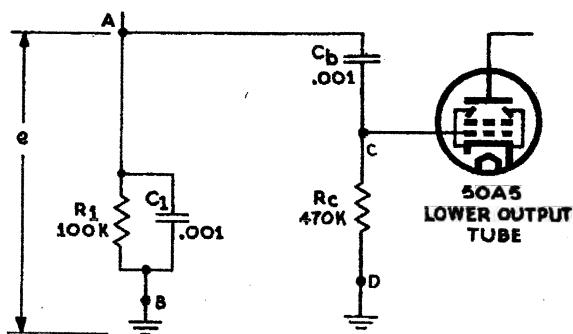

FIG. 11.—Simplified circuit of the input to the lower 50A5 output tube in Fig. 10.

combination of R_1-C_1 discriminates against high audio frequencies and the complete circuit acts together to keep the total frequency response at a constant level.

Testing Phase Inverters

In making the following tests, it is desirable to feed a weak audio signal into the grid of the audio tube with the volume control set at maximum; otherwise, tune in on a weak signal and set the volume control at maximum. In any case, *DO NOT OVERLOAD* the receiver.

One of the simplest and best methods of testing to determine whether the phase-inverter circuit is operating normally is to measure the a-c voltage at the grids of the push-pull output

the same vertical plate of the oscilloscope through another high resistance, but *of equal value to the first one used*, to the grid of the other output tube. If the phase-inverter circuit is operating normally, a straight horizontal line should be seen on the oscilloscope. This line is the resultant of the two equal but out-of-phase voltages driving the output tubes. (See Fig. 12).

If the phase-inverter circuit is not operating correctly, then a sine wave of greater or lesser amplitude, depending upon the degree of inequality of the voltages driving the output tubes, will be the resultant seen on the oscilloscope. (See Fig. 13).


Fig. 12.—When sine-wave voltages impressed on the control grid of the push-pull tubes are equal, and 180 degrees out-of-phase, they cancel and a horizontal line is seen on the oscilloscope screen.
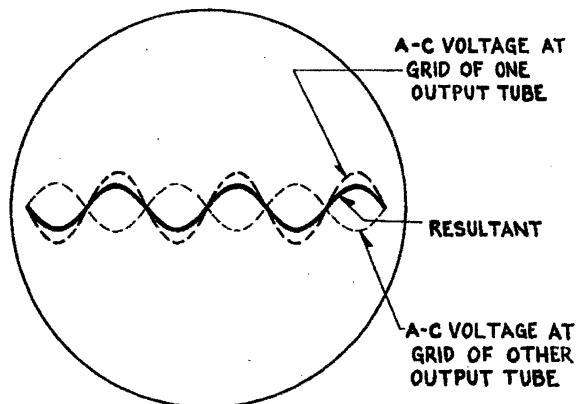

Fig. 13.—If one of the two input sine-wave voltages to the control grids of push-pull tubes, is greater than the other, then the resultant seen on the oscilloscope will be a wave.

tubes with an a-c vacuum-tube voltmeter or a cathode-ray oscillograph. The input resistance should be in the order of 200,000 ohms.

When voltages measured at the grids are of equal value, the phase-inverter circuit is working properly. If the voltages measured are unequal, then the output tubes are getting unbalanced drives from the phase inverter. Assuming that all the d-c voltages are correct, then the most probable source of trouble causing the unequal a-c voltages would probably be incorrect resistance values of the phase-inverter load.

A better method of testing these circuits is accomplished through the use of an oscilloscope. Connect a test lead from the vertical plate of the oscilloscope through a high resistance, approximately 3 megohms, to the grid of one of the output tubes. Connect another lead from

Increased Audio Power Output
From Push-Pull Operation

One of the fundamental properties relative to push-pull operation is that increased audio power output is procured as compared to that power output obtained when only one tube is used. The knowledge of increased power output is often taken for granted by many, but few actually know *how* this increase in output is obtained. Since this topic is considered to be important it is believed that a clarification of how the increased power output is obtained, is necessary.

In Fig. 14 is shown a typical push-pull circuit. The tubes are assumed to be identical in characteristics and the push-pull output transformer is center tapped, thereby allowing for equal supply voltage on both tubes. If there is

no signal input to the tubes and the proper d-c voltages are applied, then only *d-c* plate current will flow in both tubes. The flow of these d-c plate currents are indicated in Fig. 14 as straight lines. Current I_1 is that representative of tube V_1 and current I_2 is that representative of tube V_2. Since direct current flows from cathode to plate in the tubes, then currents I_1 and I_2 are seen to flow toward each other, in their own individual halves of the output transformer primary. These currents produce magnetization effects in the transformer core, but these are of opposite direction to each other. Consequently, the total magnetization effects are reduced and d-c saturation of the transformer core is likewise reduced. As stated at the beginning of this section, this reduction in d-c saturation allows greater amounts of direct current to flow through the primary and consequently greater overall audio power output can be realized.

If at any one instant of time a positive signal is assumed to exist at the input to the grid of tube V_1, then, according to one of the necessary properties of proper push-pull operation, a negative signal is applied to the grid of tube V_2. Both signals, even though opposite in phase are equal in amplitude (i.e. voltage.) This is indicated by the positive and negative half-cycle sine waves of the same amplitude, existing at the grids of tubes V_1 and V_2 respectively.

Originally, without any input signals applied to the circuit but with the proper d-c potentials applied, d-c plate current flowed in the circuit as mentioned before. Since the tubes are electrically symmetrical, the direct current flow in both tubes is the same. For the sake of argu-


Fig. 14.—In this schematic of an output push-pull stage, the output a-c and d-c plate currents are shown with respect to the input signals as well as the current in the secondary of the output transformer.

ment, the direct current is assumed to be equal to 80 ma, and this current is shown as a heavy straight line in the graph of Fig. 15. When any type of reference is made to output *signal currents*, these output currents actually consist of an a-c component of current varying about a d-c component. This d-c component is nothing more than the direct current flowing in the circuit without any signal voltages applied to the input of that circuit; consequently, this d-c output current is used as a reference level with respect to the a-c components of the current.

Effects of Input Signals

Referring back to Fig. 14, it is shown that a positive-going signal is impressed on the grid of tube V_1. This produces an a-c output current of the same frequency as the input signal. Since this input signal is positive, the plate current in the tube increases. This is illustrated in Fig. 15, in which curve No. 1 is that a-c output signal from tube V_1. Notice how this current varies above the 80-ma d-c reference level. Since only a half-cycle of the input signals on the grids are considered (for the moment), then the first half-cycles of the a-c output plate currents will likewise have to be considered as the output currents corresponding to the input signal voltages.

From Fig. 15 it is seen that the first half-cycle of a-c output current from tube V_1 (curve 1) increases the total output plate current of V_1 to 120 ma. Taking the d-c line as the reference level of the system, then the a-c plate current i_1 of tube V_1 flows in the direction, as illustrated in Fig. 14. It flows in the same direction as the direct current I_1, because of the *increase* in the total output plate current with *respect* to the d-c reference level. This a-c output current effectively increases the total output of tube V_1 by 40 ma.

Now at the same time that the grid of tube V_1 has a positive-going signal, the grid of tube V_2 has a negative-going signal. This produces an a-c output current of the same frequency as the input signal. Since both input signals to the grids are opposite in phase (180°), then both the output currents from the two tubes are also opposite in phase and of the same frequency. This negative-going input signal to tube V_2 momentarily increases the effective bias
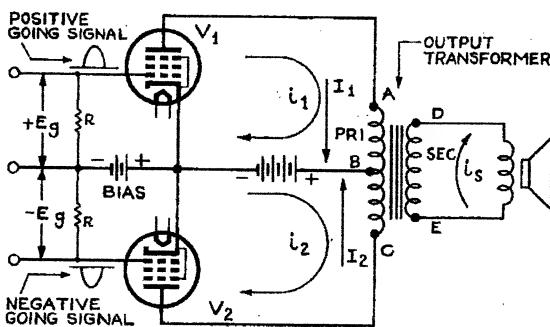
on the tube (i.e. makes the bias more negative). Consequently, the a-c output plate current from tube V_2 is decreased due to the increase in effective bias. This is illustrated in Fig. 15, where the first half-cycle of curve No. 2 is that representing the output from tube V_2. This latter output current likewise varies about the d-c reference level and for the first half cycle curve No. 2 decreases to a minimum value, which is 40 ma less than the d-c reference level. The a-c plate current i_2 of tube V_2 flows in an *opposite* direction to the d-c plate current I_2, as illustrated in Fig. 14, because of the *decrease* in output plate current with respect to the d-c reference level.
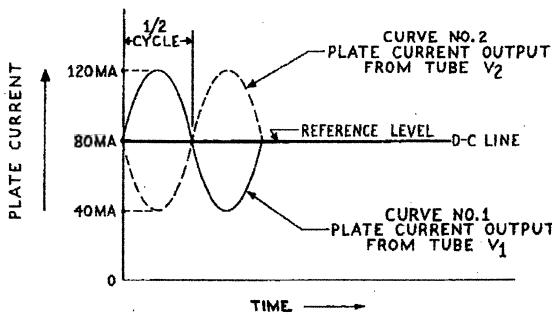

FIG. 15.—The a-c components of the plate currents of the push-pull output tubes are shown in their relationship to each other and how they are superimposed on the d-c plate current as a reference level.

In other words, what is happening in both tubes is that when the *total* plate current of one tube is increasing, the *total* plate current of the other tube is decreasing; however, when the a-c component of one tube is increasing in a positive direction, the a-c component of the other tube is increasing in a negative direction. Moreover, when the a-c component of one tube has increased to its positive maximum, the a-c component of the other tube has increased to its negative maximum. It should be remembered that these a-c plate currents are superimposed upon their direct currents and consequently the increase or decrease is primarily in reference to the d-c level.

Referring to Fig. 14 again, it will be noticed that the a-c output plate currents, i_1 and i_2, of both tubes are flowing *in the same direction* through their own individual halves of the primary winding of the output transformer. Since they flow in the same direction, their combined effects are additive and each current, i_1 and i_2, induces voltages into the secondary of

the output transformer that are in phase with each other. Since these induced voltages are in phase with each other, they produce an effective secondary current i_s that is due to the combined effects of i_1 and i_2. In other words, the secondary current i_s is greater than a secondary current produced by either i_1 or i_2 alone. This is best illustrated by the curves of Fig. 16.

Curve No. 1 is the alternating current flowing in the secondary due to the induced voltage caused by the current i_1 in the primary. Likewise, curve No. 2 is the alternating current flowing in the secondary due to the induced voltage caused by the current i_2 in the primary. In other words, current i_1 induces a voltage into the transformer secondary, as also does current i_2. Since these currents flow in the same direction in the transformer primary, the voltages that they induce into the secondary are *in phase* with each other. Consequently, these induced in-phase voltages produce *in-phase* currents in the transformer secondary, which are the current curves No. 1 and No. 2 of Fig. 16. Note how these curves are in phase with each other. (For a more detailed discussion of phase relationships see section on "Feedback.")

The currents of curves No. 1 and No. 2 are supposed to have the same amplitude (two units), because the effective values of i_1 and i_2 are equal to each other due to the electrical symmetry of the push-pull circuit. However, since it is difficult to show two curves of the same phase, the same frequency, and of equal amplitudes on one set of axes, they are shown slightly displayed from each other for purposes of illustration; actually, they should be shown as a single sine wave.

Since the individual currents flowing in the transformer secondary are in phase with each other, they are therefore said to be aiding each other; consequently, they produce a resultant
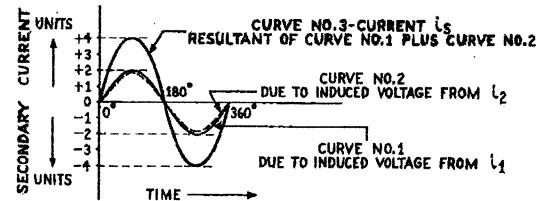

FIG. 16.—As the voltages induced in the secondary of the push-pull output transformer are in phase, the currents are also in phase and the resultant current is the sum of the two, here shown slightly displaced for the purpose of clarity.

current, i_s, in the secondary circuit that is greater than either one current alone. This latter current i_s, is the effective current flowing in the secondary circuit and it is designated as curve No. 3 in Fig. 16. Since curves No. 1 and No. 2 are in phase with each other, they are additive, and since they are equal, the final current flowing in the secondary circuit is equal to *twice* that of either individual current. Curves Nos. 1 and 2 are 2 units each at their positive and negative maximum amplitudes, and the resultant curve, No. 3, is 4 units at its positive and negative maximum amplitude, indicating a double increase in current.

If the second half-cycles of the input signals to the push-pull grids are considered, then the output a-c signal currents flowing will be represented by those second half-cycles as illustrated in Fig. 15. In this respect the a-c output plate currents of both tubes will reverse in direction (to that shown in Fig. 14) with respect to the direct current, but they will flow always in the same direction with respect to each other in the output transformer primary and consequently their combined effects will still be additive.

From the above analysis it can be concluded that the a-c plate currents in a push-pull stage flow in the same direction in the primary of the output transformer, thereby causing an increase in the total audio power output.

F-M CIRCUIT FEATURES

In general the post-war receivers intended for the reception of frequency-modulated broadcast signals are somewhat like those manufactured prior to World War II. That this should be so is not very strange inasmuch as the character of the signals remains unchanged. The operating frequencies have been increased into that region where design becomes more critical; for that matter, the same is true about service procedure. But speaking specifically about those portions of the receiver which reflect the greatest changes in thinking, these are the limiter and the discriminator.

In general, two schools of thought exist concerning the limiter and the discriminator. Perhaps this may be attributed to a patent situation; whatever it may be, it is found that some receivers make use of a limiter and one type of discriminator and other types of receivers dispense with the limiter and employ a different type of discriminator which makes this possible.

To appreciate the relationship between the limiter and the discriminator, it is necessary to understand some of the fundamental facts concerning frequency modulation. Basically the acceptance of frequency modulation for broadcasting purposes can be attributed to the advantages which this type of emission affords. First and foremost is the fact that the received signal is subject to very much less interference from noise, for the simple reason that while noise, man-made or natural, which is picked up by the receiving antenna will influence the amplitude of an amplitude-modulated wave and in that way be heard in the loudspeaker just as if it were the modulation, it will not influence the frequency of the received signal. Therefore, if a system is predicated on the transmission of intelligence by a change in frequency rather than a change in amplitude of the wave, it will not be subject to noise interference as long as changes in amplitude of the frequency-modulated wave due to the noise are prevented.

The successful accomplishment of this condition poses numerous problems to the receiver design engineer. First of all, the prevention of the acceptance of noise by a receiving antenna is almost a hopeless task. In other words, both noise and the f-m signal will find their way into the receiver. Having passed into the tuned circuit, the noise pulses will increase and decrease the amplitude of what was originally a constant value of voltage representative of the f-m signal, thus creating in effect an f-m signal which also varies in amplitude.

If such a combined f-m and amplitude varying signal finds its way into the conventional f-m discriminator, it will be rectified and both the noise as well as the intelligence will be made audible. To prevent this, a limiter stage, or perhaps two limiter stages, are utilized ahead of the conventional discriminator; these limiter stages function to remove all extraneous amplitude variations by clipping the peaks of the signal voltages. The designers of some of the modern receivers have gotten around the use of the limiter by the development of a discriminator system which is made insensitive to amplitude variations. Volume XV contains examples of both types of f-m receiver design. In order to describe these systems most effectively, it seems advisable to delve into some of the general details of f-m transmission and reception.

Differences Between A.M. and F.M.

Speaking in generalities which are sufficient for the purpose, the fundamental differences between a frequency-modulated and an amplitude-modulated wave can be expressed as follows:

1. During modulation
   a. the *amplitude-modulated* wave varies in average power in accordance with the strength of the modulating signal, but remains constant in frequency
   b. the *frequency-modulated* wave remains constant in power, but changes in frequency

2. The *frequency* of the modulating signal determines
   a. the rate of increase and decrease of power of the *amplitude-modulated wave*
   b. the rate of change in frequency of the *frequency-modulated* wave

3. The intensity of the modulating signal determines

    a. the change in average power of the *amplitude-modulated* wave

    b. the extent of the change or deviation in frequency of the *frequency-modulated* wave.

There are, of course, numerous other differences between the amplitude- and frequency-modulated waves, but their inclusion in this discussion is not necessary in order to lay the groundwork for the description of the limiters and discriminators shown in Volume XV. However, it is necessary to make some pertinent comments concerning the frequency-modulated wave.

Considering the shift or deviation in frequency created by the modulating signal, the standards set up for broadcasting technique stipulate a maximum swing or deviation of 75,000 cycles, or 75 kc, above and below the carrier for 100 percent modulation of audio signals up to and including 15,000 cycles. As is evident, f.m. provides for the transmission of a much wider band of audio frequencies than a-m transmission. To provide adequate separation between f-m stations, operating frequencies allocated to f-m stations operating within any one service area are set at least 400 kc apart, and on a nationwide basis, no two

FIG.1A

FIG. 1B

FIG.1C

FIG. 1.—Three block diagrams illustrating superheterodyne receivers: for a-m reception in Fig. 1A, and for receiving f-m signals in Figs. 1B and 1C. Note the absence of a limiter in the bottom diagram.

adjacent channels are closer than 200 kc to one another. The minimum of 200-kc adjacent channel separation is set by the maximum frequency deviation of plus and minus 75 kc for 100 percent modulation, plus 25 kc "guard" bands beyond the maximum swing in each direction.

In passing it might be well to mention that what has been stated is known as "*wide-band*" f.m. in contrast to frequency-modulation systems which employ a much reduced swing in frequency. These, however, are not employed for conventional broadcast service. For example, narrow-band f.m. limited to a swing of about 3000 cycles, is employed in amateur radio transmissions; the U. S. Navy used some transmitters with a deviation of plus and minus 15 kc; and the U. S. Army used some with a swing of about 30 to 40 kc each side of the resting frequency, with 20-kc guard bands.

The F-M Superheterodyne

Although not a necessary requirement, the f-m receiver is a superheterodyne. Up to this writing no other type of receiver capable of f-m wave reception, has been offered to the public. By this we mean that no t-r-f receiver capable of f-m reception has been marketed. The same is true of double-detection superheterodynes. Such a receiver for a-m reception, but used as a single-detection superheterodyne for f-m reception, is described elsewhere in this publication.

Being superheterodyne receivers, the f-m superheterodyne employs the usual principles of heterodyne operation in all respects; this condition is not altered by the use of the limiter and the discriminator for demodulation of f-m signals. However, in view of the nature of the frequency-modulated signal with respect to the frequency range which the radiated signal embraces, the i-f circuits are designed for acceptance of at least 75 kc each side of the quiescent or mid-frequency. This is in contrast to the usual 10-kc overall bandpass to be found in a-m receivers.

For a comparison between the a-m and f-m superheterodynes shown in Volume XV, three fundamental types are illustrated in Fig. 1 by means of block diagrams. Fig. 1A is the conventional a-m receiver of the single-detection variety; Fig. 1B is the conventional f-m re-

ceiver using the limiter followed by the conventional discriminator; and Fig. 1C is the f-m receiver which dispenses with the limiter and employs a discriminator different from that used in conjunction with the limiter.
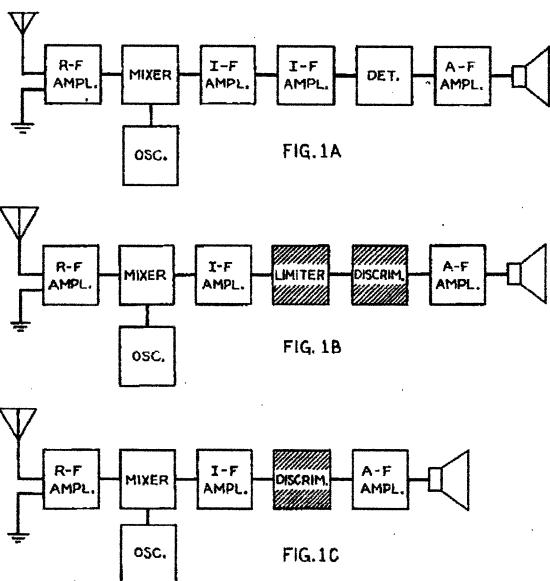
The receiver intended for the reception of frequency-modulated waves is in many respects like the receiver intended for the reception of amplitude-modulated waves.

One of the differences is found in the design of the r-f and i-f circuits. Not that the f-m receiver r-f and i-f systems look different upon paper from the representation of the a-m receiver, but the physical design of the transformers is different in order to provide the proper bandpass. This is more particularly true in the case of the i-f amplifier than in the r-f system, because the ratio of the bandwidth to the actual resonant frequency is much greater in the former than in the latter. Whereas the general run of i-f systems in the conventional broadcast type of superheterodyne used so far for a-m reception operates with an i-f peak of from 175 kc to about 465 kc and a bandwidth of approximately 10 kc, the f-m receiver employs an i-f peak of from 8 mc to perhaps 11 mc and a bandwidth of about 150 kc, or 75 kc each side of the i-f peak. In some early f-m receivers, an i-f peak of about 4 mc was used.

As a part of the i-f system of most f-m receivers, and operating at the i-f peak is a stage identified as the "limiter," which is shown in Fig. 1B. This is not entirely new to superheterodynes in that it was used in a double superheterodyne manufactured several years ago. (Rider's Volume VIII, Westinghouse page 8-11, 12). The general function of the limiter, whether a single stage or two stages are used, is to remove amplitude variations which may have developed in the carrier as the result of the addition of noise or other influences associated with the operation of the receiver ahead of the point of demodulation where the change in frequency of the carrier is converted into audio frequencies corresponding to the intelligence being transmitted. The limiter is the first difference between the a-m and the f-m receivers.

As was stated earlier, all receivers do not employ a limiter; some designers have eliminated the need for its use by the development of a new type of discriminator, another name

for the demodulator in the f-m type of superheterodyne. At any rate, the discriminator is the second difference between the a-m and the f-m receiver, that is, between the conventional a-m receiver of Fig. 1A and the conventional f-m receiver of Fig. 1B. If, however, we were comparing the a-m receiver and the f-m receiver which was designed so as not to use a limiter, then the only distinction between the two types would be the discriminator, as is seen by comparing Figs. 1A and 1C.

As an item of passing interest, the discriminator is neither new nor native only to f-m receivers. It appeared years ago in the a-m receiver equipped with automatic frequency control, which, incidentally, may again be found in f-m receivers to be made in the future. Also it is used in some f-m transmitters to keep the center frequency at the proper value.

Once past the discriminator, the receiver is identical to those already in use. Because of the higher audio range used with the frequency-modulated form of transmission, these receivers will have high-fidelity audio systems, but as far as their basic operation is concerned, one audio system is like another.

Function of R-F Tuned Circuits in F-M Receivers

As in any superheterodyne receiver, the r-f circuits used in f-m receivers have as their primary function the selection and amplification of the desired signal and the rejection of all other signals. Thus the r-f circuits must attenuate interfering signals on adjacent channels and at the same time must reduce the image response to which all superheterodynes are subject.

A feature of r-f circuits used in f-m receivers is the provision usually made in the input circuit for coupling to a balanced low-impedance transmission line, such as is used for reception at the high frequencies allocated to f-m transmissions. This feature is incorporated in the balanced-to-ground circuit shown in Fig. 2, which shows the antenna coil balance-to-ground of the Stromberg-Carlson Model 1121. In this model, a balanced-to-ground circuit in the primary winding of the f-m antenna transformer is used. The desirability of this arrangement is readily seen when one considers that interference-free reception is aided by correct design of the f-m antenna system.

The single-wire lead-in, of the type generally employed for a-m reception, is highly responsive to interference, whether natural or man-made.

A double-line feeder is generally used in f.m. to connect the receiving dipole to the antenna transformer. An interfering voltage that is induced in the double-line feeder in Fig. 2 will cause currents, as indicated by the solid arrows, to flow in the same direction through each of the two feeder wires. The current flowing through section AG of the antenna transformer primary is cancelled by the current flowing through section BG, since the fields created by these currents are equal and opposite. Consequently, the interference is not induced into the secondary of the antenna transformer.

On the other hand, a signal voltage will induce a current, as indicated by dashed arrows, which flows in the same direction through each half of the dipole. Moreover, it will be seen that this requires that this current flow in opposite directions through the feeder wires and the antenna primary. The fields due to the signal currents are therefore additive, and induce a voltage in the secondary coil, which is transferred to the first r-f tube.

Another arrangement to balance the transmission line is to return the center-tap on the primary winding of the f-m antenna transformer to ground through a wavetrap consisting of a series coil and capacitor connection. This method has an advantage in that a signal voltage is developed across this combination for the regular broadcast and short-wave bands, and while on these bands the transmission line is no longer balanced, the combination functions to pick up a signal and apply it to the input circuit. This allows the use of the same antenna for the broadcast and short-wave bands as well as the f-m band.

In addition to the advantages of r-f tuned circuits in minimizing interference and image
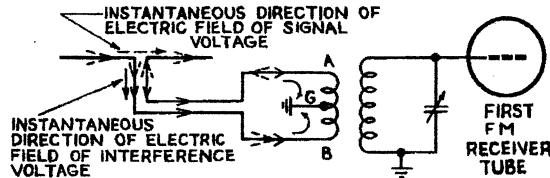

Fig. 2.—Because interfering voltages induced in the double-line feeder result in currents flowing in opposite directions in the center-tapped primary of the transformer, they cancel and thus no interfering voltage is induced in the secondary.

response, the use of an r-f stage has the very desirable effect of increasing the overall gain of the receiver and providing a higher signal-to-noise ratio. This reduction in noise level is achieved because a tube used as an r-f amplifier, has a lower inherent noise level than the same tube operated as a mixer. Thus the effectiveness of an r-f stage in reducing noise is due to the fact that the gain of the r-f stage is sufficiently great to make the signal ride above the noise level of the converter stage.

Mixer and Oscillator Functions

Operation of the oscillator and mixer stage, in an f-m receiver is not changed by the fact that f-m signals are received. The general action of these parts of the receiver conforms with those found in the conventional superheterodyne receiver and as such, the general conditions governing mixers and oscillators still apply. There are, of course, certain pertinent facts associated with performance which justify further discussion because of the servicing aspects. This is especially true of the heterodyning oscillator.

Oscillator Drift in FM Receivers

Oscillator drift is a very much more serious problem in f-m receivers than it is in conventional broadcast and short-wave receivers. It may be assumed that the carrier center-frequency remains constant in frequency. Oscillator drift will therefore manifest itself as an intermediate frequency that differs from the correct value. If the change in oscillator frequency is not too great, no significant distortion will occur in the i-f amplifier, since its passband is sufficiently wide to permit some variation.

The situation is different at the discriminator, however, for even a very small drift in oscillator frequency will result in serious audio-frequency distortion at the output of the discriminator. For example, assume an oscillator frequency of 100 mc and also that the receiver has an intermediate frequency of 10 mc. Then, if the oscillator drifts as little as 0.02%, this would correspond to a 20-kc change in oscillator frequency, and a 20,000-cycle change in the discriminator center-frequency. The discriminator is usually designed to have a bandwidth

somewhat more than about twice the deviation frequency. Due to non-linearity, the working portion of the characteristic is considerably less than this. Consequently, if the signal is not applied close to the center operating point of the discriminator characteristic, audio-frequency distortion will occur. It is, therefore, seen that oscillator drift should be reduced to a minimum.

The principal causes of oscillator drift are capacitance changes in the oscillator tube and
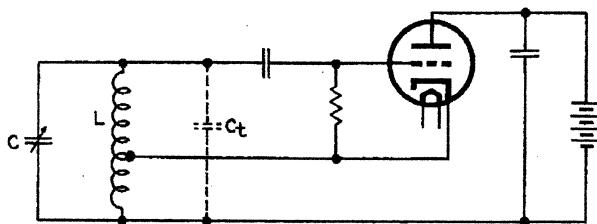
Fio. 3.—A typical Hartley oscillator circuit, the input capacitance of which is indicated by C_t.

changes in the inductance and capacitance of the tuned circuit. In well-designed receivers, the oscillator drift due to changes in the tuned circuits are not as great as those caused by capacitance changes in the oscillator tube.

An interesting method of reducing oscillator drift is employed in the Philco 46-480 receiver, by tapping-down the oscillator section of the tuned circuit. First, let us examine a typical Hartley oscillator circuit, as is illustrated in Fig. 3. The tube input capacitance C_t is shown in dashed lines. The oscillator frequency is determined by the inductance L, the tuning capacitance C, and the tube capacitance C_t. Any variation in the tube capacitance will cause a change in the oscillator frequency, and this effect will be a maximum since the tube capacitance is effectively across the entire inductance.

Now consider the circuit illustrated in Fig. 4. This circuit is identical to that of Fig. 3, except that the oscillator tube is connected across only a portion of the L-C circuit. Consequently, a change in tube capacitance causes a much smaller detuning effect, and therefore a much smaller shift in the oscillator frequency. The stability of the oscillator is accordingly greatly improved.

The minimization of oscillator drift cannot be indefinitely improved by this method however, for not only must a definite relation exist between the grid-cathode and cathode-plate

turns, but also the circuit may develop parasitic oscillations if tapped down too far.

Another method of reducing oscillator drift is exemplified in Zenith Model 12H090, Chassis 11C21. In this receiver, the oscillator is operated so that the second harmonic of the oscillator beats with the incoming signal. In other words, the oscillator frequency is one-half the sum of the signal frequency plus the intermediate frequency. This means that the oscillator can be loaded with a much higher value of lump circuit capacitance. The larger the value of the lump circuit capacitance, the less will be the effect of any small changes in capacitance due to tube capacitance or other circuit capacitance, for these latter capacitances become smaller in comparison with the former. Consequently, the frequency drift resulting from a change in the tube or circuit capacitance is very much less than if these capacitance variations had occurred when they were shunted across a smaller value of the lumped circuit capacitance that would have been used at fundamental frequency operation.

The I-F Amplifier in F-M Receivers

As far as function is concerned, the intermediate-frequency amplifier utilized in the f-m receiver performs exactly the same functions as the i-f amplifier in the a-m receiver. Modern design, however, has introduced a number of innovations which are quite interesting and in a way departures from the conventional. These shall be shown as the discussion develops, but before describing these it might be well to make several comments concerning the operating frequencies. The Radio Manufacturers Association (RMA) has standardized upon an intermediate frequency of 10.7 megacycles for all f-m receivers intended for reception of f-m signals on the band from 88 to 108 megacycles, though it is true that all receiver manufacturers
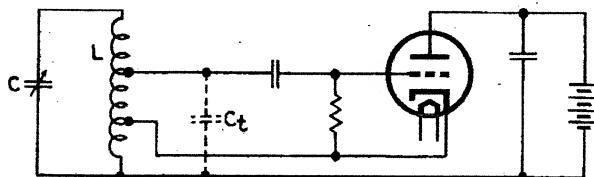

Fio. 4.—This oscillator circuit, used in the Philco Model 46-480 receiver, is the same as that in Fig. 3 except that the tube is connected to a portion of the L-C circuit.

do not adhere to this standard, many sets employing i-f systems peaked between 8.3 mc and 9.1 mc. Inasmuch as some f-m stations are still operating over the old f-m band (although conversion to the new frequencies is imperative by the end of 1947), receivers intended for the reception of these signals, as well as those radiated at the higher frequencies, employed a single i-f peak, usually that dictated by the needs of the higher carrier frequencies.

If the schematics of the various am-fm receivers throughout Volume XV be examined, it will be seen that the schematic representation and the circuits themselves are conventional. In some instances, to secure proper bandwidth, loading resistors have been used. In some systems the i-f transformers are connected in series (see Fig. 5), one set being used for the amplification of amplitude-modulated signals at a low intermediate-frequency peak and the other set being used at a high i-f peak for amplification of the f-m signal.

IF Transformers in Series

In order to save space and to avoid the necessity for switching the i-f transformers when changing between f.m. and a.m., several manufacturers, among them Philco, Stromberg-Carlson, and Zenith, are connecting the f-m and a-m intermediate-frequency transformers in series.

In Fig. 5, section *AB* represents the primary and *EF* the secondary of the frequency-modulation i-f transformer, while section *BD* is the primary and *FG* the secondary of the amplitude-modulation transformer.
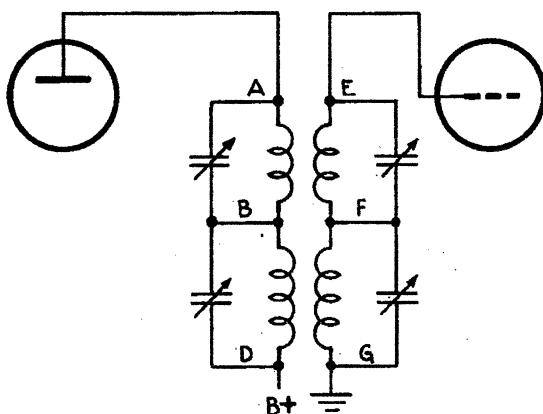

Fig. 5.—As a space-saving means, the primaries and secondaries of the i-f transformers for a-m and f-m reception, are connected in series.

At the intermediate frequency employed for f.m., the reactance of the trimmer capacitors across the a-m transformer winding is so low as to practically short-circuit these windings. Consequently, the tuned f-m circuit acts as if it were alone in the circuit. At the intermediate frequency used for a.m., the very few turns on the f-m transformer offer so low a reactance that they act as a virtual short-circuit across the f-m trimmer capacitors. Accordingly, the a-m circuit behaves as though it alone were present.

Fig. 6 on page 136 shows an exploded view of the mechanical construction of the first i-f transformer for both a-m and f-m reception, used in the Zenith Model 12H090 receiver. In this case permeability tuning is used; it is the fixed capacitance which is in parallel with the a-m transformer windings that short-circuits these windings at the intermediate frequency used for f-m.

The independence of the f-m and a-m intermediate-frequency transformers in operation does not signify that they can be aligned in any arbitrary sequence. Quite the contrary is the case. For proper alignment, *it is absolutely essential that the a-m transformers be aligned before the f-m transformers.*

Suppose, for example, that the f-m circuits had first been aligned. Then, in the process of aligning the a-m circuits, the f-m circuits would become misaligned. For, although it is true that the a-m trimmer capacitors have an extremely small reactance at the intermediate frequency used for f-m, this reactance is not absolutely zero. Since it requires a very minute change in impedance at these high frequencies to cause an appreciable change in circuit frequency, the f-m circuits become detuned in the process of a-m alignment.

On the other hand, suppose that the a-m circuits had first been aligned. The subsequent alignment of the f-m circuits does not cause any significant detuning of the a-m circuits, because the f-m circuit impedance is negligible at the a-m intermediate frequency.

The Limiter

The final i-f stage in some f-m receivers is known as the *limiter;* for that matter some f-m receivers make use of two limiter stages in cascade, in which case they are the third and

fourth i-f stages. (See Scott Model 800-B, Volume XV, page 15-31, 32). The primary purpose of the limiter in the f-m receiver is to remove whatever amplitude variations that may have developed on the signal during its entry into, or passage through the receiver, due to noise or the presence of an interfering signal. The removal of such amplitude variations from the signal is very important, because the conventional discriminator is responsive to amplitude as well as frequency variations, and if either an interfering signal or noise is the cause of such variations, both will appear in the output of the receiver.

In this connection it is important to reiterate that the f-m receiver which does not use a limiter accomplishes the removal of amplitude variations, or freedom from response to amplitude variations in the discriminator, by the use of special circuits ahead of or in the discriminator stage.

The operation of the limiter is comparatively simple, the circuit arrangement being a system wherein the vacuum tube overloads quite readily and in so doing cannot follow faithfully the varying positive and negative peaks of the input signal, delivering to the load in the plate circuit a signal voltage which is constant in amplitude for all signal input voltages greater than a threshold value. Thus, "limiting" of the signal output voltage takes place, with no restriction of frequency variations.

Single-stage limiters are generally of similar type but with minor differences in circuit arrangement, as shown in Figs. 6A, 6B, and 6C. The major difference is found in the wiring of the grid capacitor and grid resistor, C and R respectively. In some instances, a parasitic resistor of very small value (from 25 to 50 ohms), may be found in the grid circuit at the control grid of the tube.

In order that the proper operating state exist in the limiter, certain conditions must be met. In the first place, the tube used must be of the sharp-cutoff type, like the 6SJ7 or the 6AC7, with comparatively low screen and plate voltages and no fixed control-grid bias. Under such conditions, input signals of comparatively small amplitude (determined by design) will swing the grid voltage into that zone which will cause plate-current cutoff on the negative peaks; rectification in the grid circuit on positive peaks restricts the rise in plate current, thus effecting clipping of both negative and positive peaks. Operating plate and screen voltages as found in receivers, are within a 45-to 90-volt range.

Graphical portrayal of the operation of the single-stage limiter is illustrated in Fig. 7. The actual circuit used for the development of these data are shown in Fig. 6B, which, in operation would be like Figs. 6A or 6C. The values used for the components and the operating potentials conformed with those found in commercial receivers, so that the description can be said to apply to all the single-stage limiter systems shown in Volume XV, for that matter even in earlier volumes.

The values of plate current given are those that were actually present with different grid voltages. Note that the plate current reaches cutoff when the control-grid voltage becomes


FIG. 6A

FIG. 6B

FIG. 6C
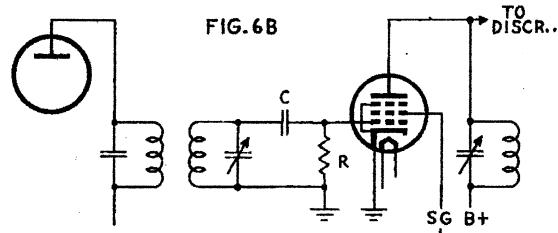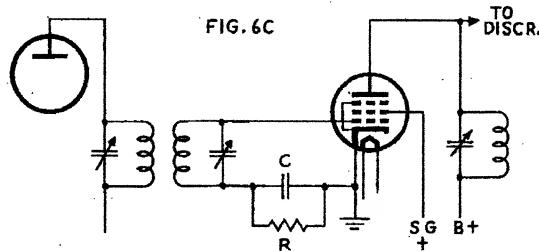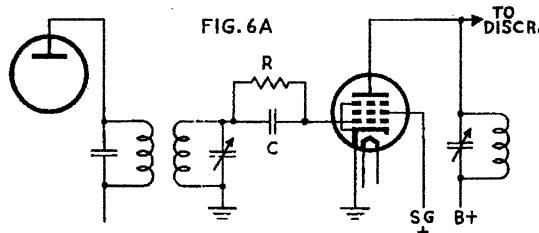
Fig. 6.—Three different types of single-stage limiters. The major differences are to be found in the location of the grid capacitor and grid resistor, C and R respectively.

approximately 8 volts negative. This means that any signal with a negative peak voltage which together with the d-c bias exceeds 8 volts, will place a total negative bias on the grid sufficient to drive the plate current to zero. Consequently, since no plate current flows when the negative grid voltage exceeds 8 volts, the negative peak of the wave is cut off in the plate circuit. This clipping of the negative peak is illustrated in the diagram Fig. 7.

This does not mean that the applied voltage must exceed 8 volts on the negative peak to restrict the rise in plate current, though such would be the case if the effective grid bias, under operating conditions, were zero. Actually when a signal voltage is applied, the control grid draws current over a portion of the positive half cycle as seen in Fig. 7. This grid current charges the capacitor C over the peak portion of the positive input signal and during the rest of the swing of the input signal the capacitor discharges through resistor R. This discharge through R is in such a direction as to put a negative bias on the control grid of the tube. This negative bias increases to a peak according to the maximum amount of discharge of C, but the increase in bias will *never* exceed the amount of voltage at the positive peak of the input signal.

Since the time constant of the combination R and C is long compared with the time interval of a single cycle of the signal, the capacitor does not discharge much and it keeps a large portion of its charge during the time when the grid is negative with respect to the cathode. Therefore, most of the negative voltage developed over the positive peak is retained. This results in a shift in the operating bias of the
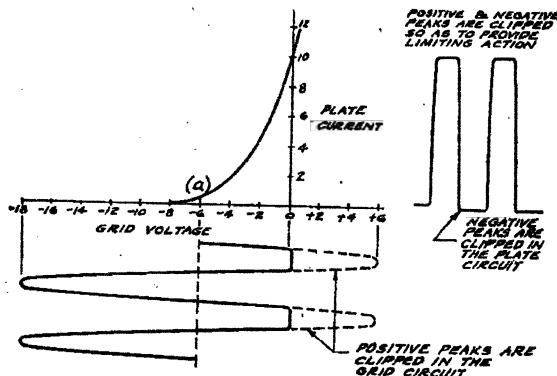
grid from zero to some point, such as (a) in Fig. 7, which depends upon the magnitude of the applied signal. Over the negative half-cycle, therefore, any signal peaks higher in voltage than the difference between the bias at point (a) and cutoff (—8 volts) are clipped in the plate circuit.

Clipping of the positive peaks takes place in the grid circuit. At the start of the positive swing when the grid draws current, the tuned circuit becomes heavily loaded with the result that the signal peak becomes flattened. Moreover, the positive signal voltage does not drive the grid several volts positive, as might be assumed when the input signal is strong. In fact, the grid becomes only slightly positive because of the residual grid bias. The only reason that practically all of the positive half-cycle is not clipped is because at each instant over the positive half-cycle, when grid current is produced, a corresponding negative charge is developed on C (in Fig. 6) because of the grid current. This serves to provide increasing negative control-grid bias, which permits a further excursion of the positive half-cycle. If the negative voltage thus developed were exactly equal to the positive peak voltage, no clipping would result. Actually, the operating bias thus developed is less; consequently, clipping does occur. The portion of the positive half-cycle which is thus clipped, is shown in dotted lines in Fig. 7.

Thus it can be said that the plate current varies between two limits: that which corresponds to plate-current cutoff and that which corresponds to the slightly positive grid. This is so regardless of the variations of the signal-voltage levels in the input circuit, assuming, of course, that the input signal voltage is sufficient to drive the tube to cutoff. If the input signal voltage is less than this amount, the tube will function as a conventional amplifier without displaying any limiting action.

The Two Stage Limiter

In general, the performance of a two-stage limiter is like that of the single stage, in fact, the circuits of the two-stage limiters to be found in Volume XV are like the single stage shown in Fig. 6B, except that two stages are cascaded. By cascading two such stages, the action of one supplements the action of the


FIG. 7.—The plate current-grid voltage curve obtained with the single-stage limiter circuit of Fig. 6B. 90 volts were applied to both the plate and screen grid.

other; whatever amplitude variations may have passed the first stage because of inadequate signal swing, will have been amplified by the tube to an extent permitted by its operating potentials, and then clipped by the succeeding stage. At the input of the second stage, the signal which was too weak to actuate the first limiter will have reached such proportions as to cause the required limiting in the second stage. Moreover, with two limiter stages in a receiver, it is possible to adjust the relative operating conditions so that the best possible performance is obtained from the second stage.

The coupling between the two limiters can be either resistive-capacitive and capable of passing the full range of intermediate frequencies encountered in the circuit, or it may be the conventional i-f transformer utilized in the receiver. The latter, of course, is preferred inasmuch as it aids the attainment of greater adjacent channel selectivity. In fact, the f-m receivers using two-stage limiters and shown in Volume XV employ tuned circuits for interstage coupling.

Limiter Action Summarized

In summarizing the action of the limiters in the f-m receiver, it is possible to view them as Class C amplifiers, wherein the output is controlled both on the negative and positive peaks by the operating potentials. These in turn are so chosen that when the threshold value of voltage is exceeded, there is no further increase in output; and the output can be less than the input. Therefore the presence of adequate gain ahead of the limiter can be readily appreciated. It is understandable that the objective in limiting is to saturate the tube with the lowest possible signal input voltage to the receiver. This can be done only when ample gain exists in the receiver between the antenna and the input circuit to the limiter.

Since the rectified grid current developed in the grid circuit of the first limiter tube varies in accordance with the input signal voltage, the rectified grid voltage developed across the grid resistor R is used as an automatic bias for automatic volume control. This control voltage may be applied to the r-f and i-f tubes, as may be desired. The rectified grid current developed in the second limiter stage, however, is unsuited for this purpose because while it may be

proportional to the input signal voltage to that limiter stage, it is not proportional to the changes in signal input into the receiver. Consequently, it cannot be used as a means of increasing the gain when the signal input to the receiver is low and to decrease gain when the signal input to the receiver is high.

It is conceivable that during the life of this text, other types of limiters will make their appearance. For instance, a two-stage limiter need not be of the kind which has been discussed; instead it can be arranged so that the first stage clips the negative peaks by means of plate-current cutoff and the positive signal peaks are amplified without limiting, then clipped in the second stage. In general, however, the procedures for limiting in the receivers produced during 1947 will more than likely be along the lines that have been described.

The Discriminator or Frequency Demodulator

The major difference between an f-m and an a-m receiver is found in the means employed to develop the intelligence from the modulated carrier after suitable amplification. In the a-m receiver this is done by converting the amplitude modulation into amplitude variations of tube plate currents, which, when finally applied to a loudspeaker, become the transmitted intelligence. In the f-m receiver, the demodulation procedure used to develop the transmitted intelligence is the conversion of frequency variations into amplitude variations of the tube plate currents. When finally applied to the speaker, the intelligence transmitted is made audible.

Numerous types of f-m demodulating processes are available, but only a few are actually in use in current receivers and these will be described. Strangely enough the most commonplace of these discriminators is not new; for that matter, it was used even before f-m receivers made their appearance on the public market. The automatic-frequency-controlled type of receiver of yesteryear made use of the discriminator as a means of developing control voltages when the heterodyning oscillator drifted off frequency; these control voltages then being used to restore again the oscillator to its correct frequency by means of a frequency-control system, the "reactance modu-

lator." More than likely these systems in time will again become popular. By no means have they been forgotten completely, since they are used in what is known as the "direct" type of f-m transmitters.

Since two types of discriminators are in general use, it might be well to identify them by their names. The most popular today is the Foster-Seeley, also known as the "center-tuned" or "phase" discriminator, which is used in those receivers which also make use of limiters. Then there is the "ratio detector," which while in some respects is like the other, does contain an important difference and is the one used in numerous receivers which do not employ limiters.

Speaking in generalities, f-m discriminators convert frequency variations into amplitude variations by the combined action of a number of conditions:

1. The phenomenon that the circulating current in a resonant circuit may be in phase with the induced voltage which caused that current to flow; it may lag that voltage or lead that voltage, the conditions being determined by the relationship between the frequency of the applied voltage and the resonant frequency of the circuit.
2. The use of a center-tapped resonant circuit wherein the voltage developed by the circulating current is equally divided for application along two paths.
3. The vector addition of a fixed phase reference signal voltage to two other signal voltages which differ in phase with respect to the reference voltage.
4. The phenomenon that the resultant amplitude of two voltages of like frequency is determined by the relative phases.
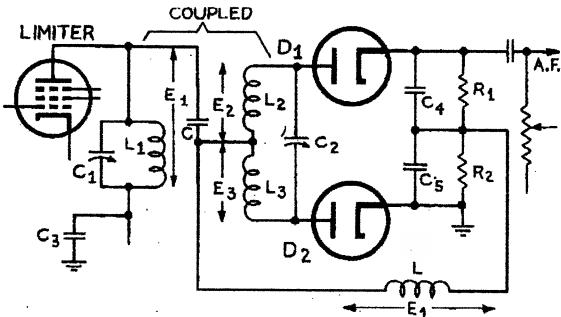
Fig. 8.—The basic discriminator circuit. Note that the transformer secondary, L_t and L_s is center-tapped and resonated to the i-f peak by the capacitor C_t.

5. The application of these combined signal voltages to a differentially connected rectifier system, wherein the output will be proportional to the difference between the amplitudes of the signal voltages applied to the two diodes.

To demonstrate the application of these phenomena, consider a typical discriminator circuit, such as is used in the f-m receiver which utilizes one or more limiters and which appears in Fig. 8. The discriminator used in the limiterless receiver resembles the one shown, but it contains a major difference, which will be discussed in detail later.

Referring to the standard f-m demodulator of Fig. 8, certain pertinent conditions must be noted. There is nothing special about the limiter plate circuit or the tuned i-f transformer primary connected therein. Capacitor C_i resonates with L_i to the i-f peak, and E_i designates the signal voltage developed across this winding.

The I-F Transformer

Examining the secondary of this i-f transformer, we note certain significant details. First, it is shown as consisting of two windings L_s and L_t in series, resonated to the i-f peak by means of C_s. The fact that the secondary is shown made up of two windings is in itself not significant; it could just as readily be a single winding tapped at the center, which would be the same thing. It is more convenient, however, to show it as two windings because of what will follow. The center tap on the secondary winding is shown connected to a coupling capacitor C and also to an r-f choke L.

Associated with the two circuits and the r-f choke L are three voltage designations E_s, E_s, and E_i respectively, the latter being identical to the E_i which is shown associated with the i-f transformer primary. To explain these designations, it is necessary to speak about the coupling between the primary and secondary circuits of this transformer, as well as what happens in a transformer when the secondary is tapped at the midpoint.

Two methods of coupling the signal from the primary to the secondary circuit are used in this system. The resonant primary is inductively coupled to the resonant secondary winding; at the same time the signal voltage E_i across the primary is fed to the r-f winding L

via the coupling capacitor C. If the circuit of C, L and C_5 is traced, it will be seen that L is in shunt with the tuned primary, the latter being grounded through C_5. Neither C, L, C_4 or C_5 are of such magnitude as to alter the resonant conditions of C_1 and L_1, the resonant primary. Thus we can set up immediately the condition that whatever signal voltage exists across C_1-L_1, the same signal voltage with respect to magnitude and phase exists across L. The direct connection between the coupling capacitor C and the mid-point of the secondary winding is of no consequence with respect to the signal transfer between the primary and the secondary circuits; it happens to be the common junction between the means of feeding the signal to the choke L, and the point to which the choke L must be connected so as to complete the differential rectifier circuit.

Thus the secondary system receives signal voltages in two ways; the resonant secondary receives its signal voltage by inductive coupling, and the r-f choke derives its signal voltage by means of direct coupling through the fixed capacitor C. Returning to the two coils which comprise the secondary winding and the associated signal voltage designations, the latter come about in the following manner. When a winding is tapped at the midpoint and a voltage is induced in that winding by means of a varying magnetic field, the total voltage developed across the entire winding divides between the two halves. This is readily evident when it is realized that half the total number of turns exists between the center tap and one end, and between the center tap and the other end. So, whatever is the nature of the signal voltage which will be developed across the tuned secondary circuit C_2-L_2-L_3 it is possible to show this voltage divided into two parts: that across each half of the winding. These are designated as E_2 and E_3. . . . So much for the elements to the left of the diodes; more details will follow later.

The Diode Circuit

Now examine the connections to the diode and the load circuit associated with the diode. Suppose for the moment we assume the existence of a signal voltage E_1 across L; likewise signal voltages E_2 and E_3. In simplified form, the complete discriminator system of Fig. 8 can be reproduced as shown in Fig. 9. Since

the output of the diodes is unidirectional current, it is possible to assign polarity signs to the two load resistors to indicate the direction of the current flow. When this is done, the nature of the circuit becomes evident.

First, whatever the character of the signal voltages E_1, E_2 and E_3, it is readily seen that signal voltage E_1 is common to both diodes; also that the voltage active on diode D_1 is the combination of E_1 and E_2 and the voltage applied to diode D_2 is the combination of E_1 and E_3. Moreover the connection of the diodes is such that whatever rectified voltage develops across the load resistors R_1 and R_2, the voltages will at all times oppose each other, and if any output will develop, it will be the differential between the voltage across R_1 and the voltage across R_2. Whether or not any differential voltage will exist greater than zero, will obviously depend upon the values of E_1+E_2 and E_1+E_3. . . . It is readily seen that if E_1+E_2 is greater than E_1+E_3, then more current will flow through R_1 than through R_2; the rectified voltage across R_1 will be greater than the rectified voltage across R_2 and the output voltage will be the difference between the voltage drop across R_1 and R_2 and the polarity of this voltage will make point X positive with respect to point Y.

If on the other hand, the reverse is true, namely, that E_1+E_2 is less than E_1+E_3, then the rectified current through R_2 will be greater than through R_1; the rectified voltage across
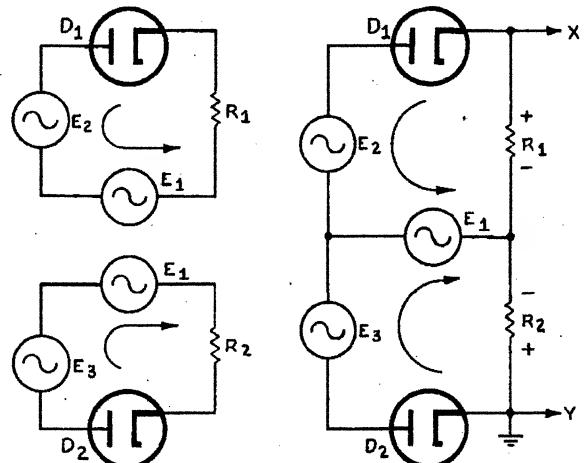

Fig. 9.—On the left are the simplified circuits of the two diodes of Fig. 8 with the voltages developed across the coils. On the right these two circuits and sets of voltages have been combined, showing the polarity of the voltage drops in the load resistors.

R_s will exceed the rectified voltage across R_i and the differential voltage will be the difference between the voltage drop across R_s and R_i, in which case point X will be negative with respect to point Y. Based upon the action described, we can identify this system as being a differentially connected diode circuit. With respect to the polarities mentioned, their significance in the operation of the f-m discriminator is not that a d-c voltage of a certain polarity will be available in the output circuit, but that the output voltage can swing negative and positive from a zero value.

Now we can return to Fig. 8 and establish the operation of the tuned circuit with respect to various conditions of resonance, then show the manner in which the various signal voltages present in the discriminator system combine to enable the conversion of a change in frequency into a change in amplitude. First of all, we shall repeat that the signal voltage E_i present across the tuned primary in the limiter plate circuit also is present across the r-f choke L, both in magnitude and phase. This is quite important to the operation of the whole system, and must be borne in mind when we consider certain phase relationships between the signal currents and voltages in C_i-L_i and in the tuned secondary circuit.

I-F Transformer Action

Let us now examine what happens in the tuned i-f transformer which couples the limiter to the discriminator diodes. For that matter, the conditions to be mentioned are not native to such i-f transformers only; they are applicable to any tuned transformer. For a start we shall assume that the primary and secondary are resonated to the same frequency and a signal voltage exists across the tuned primary. Also for the purposes of this discusssion we can forget the existence of the diodes and the load circuit, so that in effect we have the equivalent of Fig. 10 as the circuit to discuss.
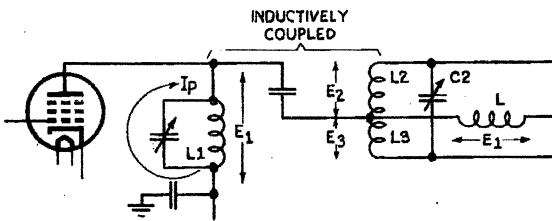

Fig. 10.—Simplification of Fig. 8, omitting the diodes and loads.

Let us start with a signal voltage E_i across the winding L_i. This voltage will cause a curcent I_p to flow through the winding, but in accordance with the basic law of current and voltage phase relationship in an inductance, the current will lag the voltage by 90 degrees. Now it is the rise and fall of this primary current which creates a varying magnetic field around the coil and cuts the turns of the secondary winding, inducing a voltage in the secondary. This voltage is at every instant proportional to the rate of change of the lines of force, being maximum when the field around L_i is changing most rapidly, that is, when the primary current is passing through zero. In turn, when the primary current is passing through its maximum, the lines of force are passing through maximum and the rate of change is zero, so that the induced voltage is zero. In the light of this condition we can say that the *induced* voltage is 90 degrees behind the inducing field, and since the inducing field is in phase with the primary current—which current is 90 degrees behind the applied voltage E_i—the induced voltage in the secondary is 180 degrees out of phase with the primary voltage. This is a fundamental condition in all transformers.

It is important at this stage to understand a very significant condition. The voltage induced in the secondary circuit by the lines of force created by the primary current is *not* the reactive voltage present in the circuit. Instead, the *induced* voltage is simply a driving voltage which causes a current I_s to flow in the tuned secondary circuit, just as if a voltage equal to the *induced* voltage were placed in series with the coil in the tuned circuit. It is this induced current which develops the *reactive* voltage across the respective L and C elements of the circuit.

What Happens at Resonance

Let us now investigate what happens during various conditions of resonance. Assume that the frequency of the applied voltage is the same as the frequency to which the secondary circuit consisting of L_s, L_s, and C_s, are tuned. When a circuit is resonant its inductive and capacitive reactances are equal and offset each other, so that the circuit behaves like a resistance. Now, in a resistive circuit the current

and voltage are in phase so that the induced current I_s caused to flow through the secondary circuit will be in phase with the induced voltage. If we now recall that the primary signal voltage E_1 is 180 degrees out of phase with the induced (series) voltage E_s in the secondary and that the primary voltage appears across L with no change in phase or magnitude, then it is possible to illustrate this much of the voltage relationships in the discriminator circuit, as appearing in Fig. 11A. Since the secondary circuit is resonant, the induced current I_s is in phase with the induced voltage and this too is shown in Fig. 11A.

Now, whatever the actual value of the *reactive* voltage developed across the series combination of L_2 and L_3 by the induced current I_s, that voltage, in accordance with the basic laws of alternating-current flow through an inductance, will be 90 degrees ahead of the current through the inductance. This is shown also in Fig. 11A as E_2,E_3. Inasmuch as a vectorial representation of this condition is easier to see, we translate the contents of Fig. 11A

into the vectorial representation in Fig. 11B. The length and direction of the lines show the magnitude and phase of the respective items. We have chosen E_1, the primary voltage, as the reference voltage and show it in the 0° angular position. Differences in phase are considered in a counterclockwise direction. As is evident, the 180-degree phase difference between E_1 and E_s is indicated by showing E_s in a direction exactly opposite to that of E_1. Since the induced current I_s is in phase with the induced voltage, it is shown in the same direction as E_s. However, as E_s is greater in amplitude than I_s as seen in Fig. 11A, then E_s is shown to be greater in magnitude than I_s in Figs. 11B and 11C. Inasmuch as the reactive voltage across L_2 and L_3 leads the induced current I_s (the induced current lags the reactive voltage), the vector for the reactive voltage E_2,E_3 is shown 90 degrees ahead of the induced current I_s.

The very important detail which determines the operation of the discriminator also is shown in Figs. 11A and 11B, this being that the phase difference between the reactive voltage across

E_1 = PRIMARY SIGNAL VOLTAGE
ACROSS R-F WINDING L
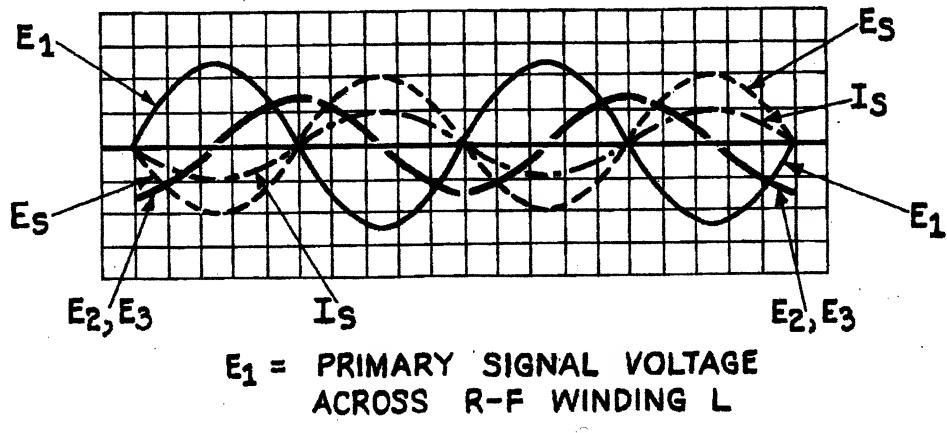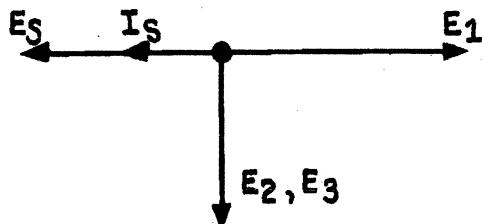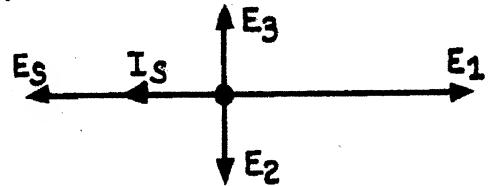
FIG. 11 A

FIG. 11 B

FIG. 11 C

Fig. 11.—Since the secondary circuit is resonant and purely resistive, the current I_s flowing in the secondary circuit of the discriminator is in phase with the induced voltage E_s. This in-phase relationship is shown graphically in Fig. 11A and vectorially in Figs. 11B and 11C with other discriminator voltages.

the coil in the tuned circuit and the voltage existing across the r-f winding L is 90 degrees. This is very important as you will soon see, for it is this phase displacement which changes under conditions of resonance and results in the uneven values of signal voltage applied to the two diodes. To appreciate this better, we show a revision of Fig. 11B in Fig. 11C. What we have done is to put on paper the fact that the coil in the tuned circuit is center-tapped, so that the total reactive voltage divides into two parts, each being equal to half of the whole. This was done by redrawing the vector E_2, E_3 so that E_2 is displaced by 90 degrees from E_1, and so is E_3, the amplitudes of each being half of the previously combined E_2, E_3. This is permissible for although the total reactive voltage across the tuned winding is 90 degrees ahead of the induced current I_S, the voltage E_2 is 180 degrees out of phase with the voltage E_3, both being displaced by 90 degrees from the reference voltage E_1. That is, voltage E_3 is leading voltage E_1 by 90 degrees and voltage E_2 is lagging voltage E_1 by 90 degrees.

Having established the phase relationships between the three signal voltages in the discriminator circuit, we can establish the values applied to the two diodes. If you will glance again at Fig. 9, you will see that diode D_1 receives the sum of E_1 plus E_2, and that diode D_2 receives the sum of E_1 plus E_3. However, since these voltages have both direction and magnitude, simple arithmetical addition does not give the answer. Graphical addition is necessary in the waveform types of representations and vectorial addition in the vectorial representation. The former types are shown in Figs. 12A,-B, and -C and the latter type in Fig. 13. In Fig. 12A, we have shown the 180-degree phase difference between E_2 and E_3, and the 90-degree phase difference between E_2 and E_1, and between E_3 and E_1. Voltage E_2 is shown as equal to E_3, because the total reactive voltage across the coil divides equally each side of the center tap. In Fig. 12B, we show voltage E_2 and E_1 in the same relation as in Fig. 12A, and also the graphical sum of these two voltages, which takes into account both direction and magnitude. The same was done in Fig. 12C, except that in this case voltages E_3 and E_1 are involved. With the waveforms shown in Figs. 12A, 12B, and 12C drawn to the same scale, it is evident that the combination of E_2 and E_1, and the combination of E_3 and E_1, results in like voltages being applied to the two diodes.

In the vectorial representation, shown in Fig. 13, the parallelograms are completed corre-
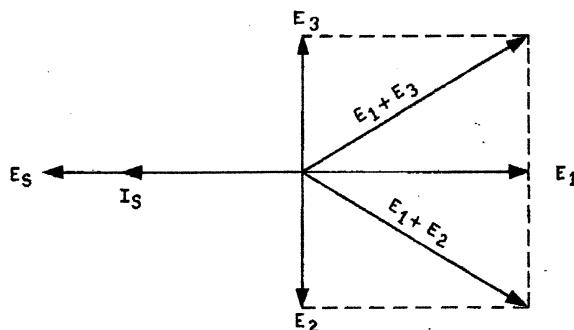

Fig. 13.—Vectorial representation of the three voltages of Fig. 12, wherein E_2 and E_3 are 180 degrees out of phase and each is 90 degrees out of phase E_1.


FIG. 12A
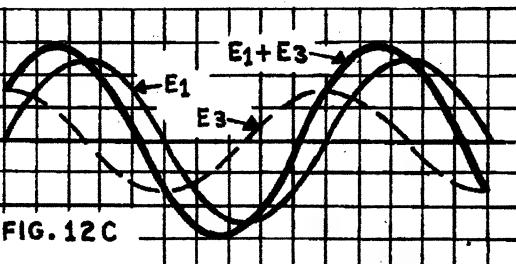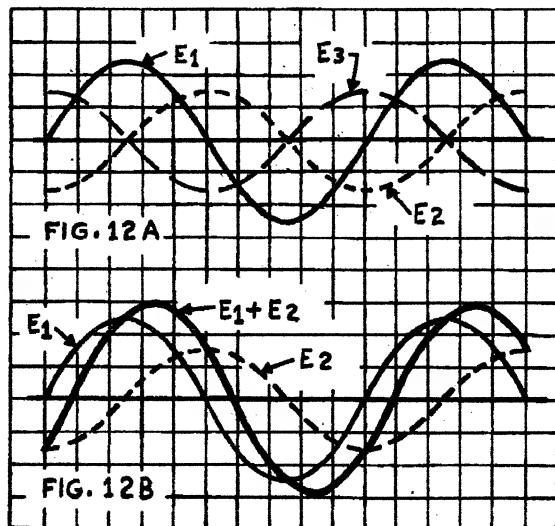
FIG. 12B

FIG. 12C

Fig. 12.—The voltages E_2 and E_3, across each half of the discriminator secondary are 180 degrees out of phase with each other; these are shown graphically in Fig. 12A with the primary voltage E_1 as reference. Figs. 12B and 12C show the addition of these voltages as applied across the diodes at the resonant intermediate frequency of the discriminator transformer.

sponding to the respective amplitudes and the diagonals indicate the resultant. As can be readily seen, the lengths of the two diagonals E_1+E_2 and E_1+E_3 are the same, hence the signal voltages they represent are the same.

With equal values of voltage applied to the two diodes, equal values of rectified current will flow through the diode load resistors and equal values of charge will be applied to the two capacitors C_4 and C_5. With equal voltages developed across the diode loads and these being in opposition to each other, the net difference is zero and no audio voltage will appear across the output circuit.

What Happens When the I.F. Is High

Let us now investigate the conditions which exist when the i-f signal is tuned *above* the frequency of the resonant circuit. This is the same as saying that the discriminator circuit is tuned below the intermediate frequency. The circuit is the same as before; the items to be considered are the respective voltages and currents and their phase. Since non-resonant conditions do not alter the fundamental rules, the action of the primary circuit remains as before. The same is true of the fact that the primary signal voltage still exists across the r-f winding L in the discriminator circuit. The induced voltage E_s in the secondary remains 180 degrees out of phase with the primary signal for this too is a fundamental condition which is not altered by resonance conditions. However, the relationship between the induced voltage E_s and the current I_s, which it causes to appear in the secondary circuit, is affected by the state of resonance, and in so doing alters related conditions.

When the applied frequency is higher than the resonant frequency, the reactance of the coil, even though it is fixed in inductance, becomes greater than the reactance of the capacitor. This is in accordance with the fundamental law that inductive reactance varies in proportion to frequency and capacitive reactance varies inversely with frequency. Accordingly, a portion of the inductive reactance will offset the capacitance reactance, but a certain amount of inductive reactance will remain to display a control on the induced current. In other words, the circuit as a whole now appears as an inductance and resistance in series, rather than as a resistance alone, which is the case at resonance.

Under the circumstances the induced current I_s will no longer be in phase with the induced voltage E_s, but rather will lag this voltage by a certain amount, depending upon the extent to which the i-f signal is off-resonance. Suppose for the sake of argument, we say that the deviation from resonance is such that the amount of inductive reactance remaining is sufficient to cause a lag of 45 degrees between the induced current I_s and the induced voltage E_s.

Now regardless of the phase relationship between the induced current and induced volt-
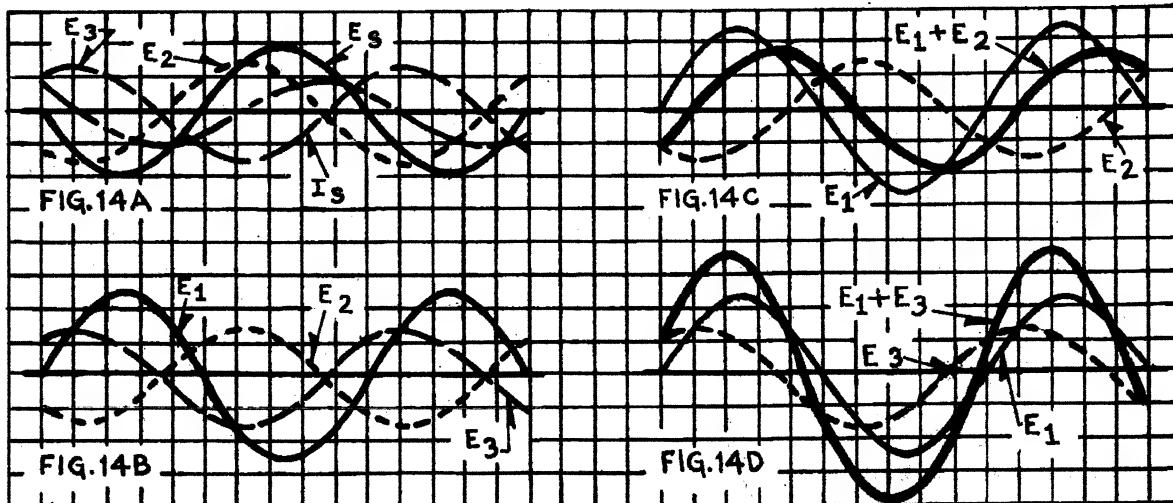
FIG. 14.—The relationship between the curves of the induced voltage and current, E_s and I_s, and the secondary voltages, E_2 and E_3, when the frequency of the input signal is higher than the resonant frequency of the secondary tuned circuit. Note that $E_1 + E_2$ is less than $E_1 + E_3$.

age, the reactive voltage developed across the coil in the tuned circuit by the induced current will still lead the induced current by 90 degrees. However, since the induced voltage E_s is 180 degrees out of phase with the signal voltage across winding L, and the induced current lags this induced voltage by 45 degrees, the reactive voltage across the coil in the tuned circuit, no longer will be displaced 90 degrees
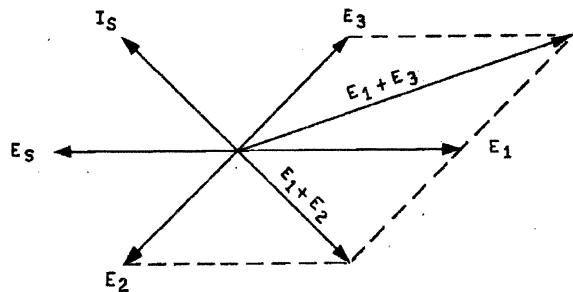

FIG. 15.—Vectorial representation of the current and voltages of Fig. 14, wherein the current I_s lags the voltage E_s by 45 degrees.

from the primary voltage E_1 existing across winding L. If for the moment we forget that the tuned secondary coil is center-tapped, the 45-degree lag between the induced current I_s and the induced voltage E_s will make the phase difference between the reactive voltage E_2, E_3 135 degrees behind the primary voltage E_1. Because of the center tap on L_2-L_3, and its junction with L, we can show the two voltages

E_2 and E_3 as being 180 degrees apart, which will make E_3 45 degrees out of phase with E_1, and E_2 135 degrees out of phase with E_1. This is shown in Figs. 14A,-B,-C, and -D and vectorially, in Fig. 15.

What Happens When the I.F. Is Low

The conditions existing when the intermediate frequency is *lower* than the resonant frequency in the discriminator are, in the light of what has been previously described, comparatively simple to understand. In this condition of resonance, the first change from the previously described conditions is that since the frequency of the signal is lower than the resonant frequency of the circuit, the reactance of the tuning capacitor C_2 predominates in the circuit, since at the frequency involved, it exceeds the reactance of the coil. Therefore, the circuit as a whole behaves capacitively, and in accordance with the fundamental law of behaviour of current and voltage in a capacitive circuit, the current will lead the voltage. This means that the current I_s, caused to flow by the induced voltage E_s, will *lead* that voltage by a certain amount determined by the degree off-resonance. We have set this as being 45 degrees so as to conform with the phase difference when the frequency of the applied
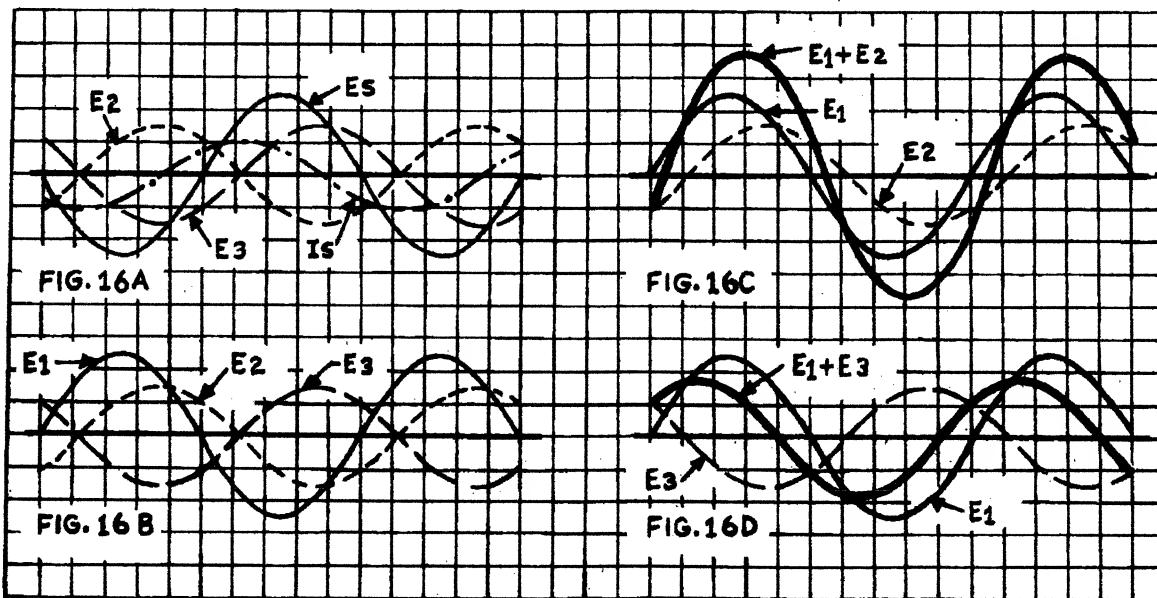

FIG. 16.—The relationship of the same voltages and current shown in Fig. 14, but in this case the frequency of the input signal is lower than the resonant frequency of the secondary tuned circuit. Note that here $E_1 + E_2$ is greater than $E_1 + E_3$.

signal is higher than the resonant frequency of the discriminator secondary circuit.

Once more the reactive voltage developed across the coil will lead the induced current I_s by 90 degrees, and since I_s leads E_s, the reactive voltages E_2 and E_3 will be displaced by 45 and 135 degrees respectively. These conditions are shown in Figs. 16A,-B,-C, and-D, and vectorially, in Fig. 17. As can be readily seen, the signal voltages consisting of $E_1 + E_2$, and $E_1 + E_3$ are unequal; hence, unequal values of signal voltage will be applied to the two diodes;
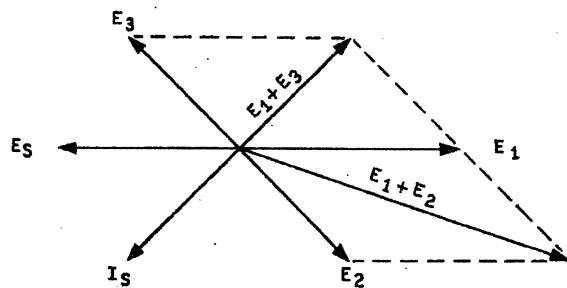

Fig. 17.—Vectorial representation of the current and voltages of Fig. 16, wherein the current I_s leads the voltage E_s by 45 degrees.

the rectified voltages developed across the diode loads will be unequal, and, in the case cited, the output voltage will be negative with respect to ground.

Summarizing the action described, you can readily see that if a varying frequency signal, one which varies in frequency around a mean, is applied to the discriminator network—provided, of course, that the range of frequencies covered is not beyond the acceptance bandwidth of the discriminator transformer—a signal which changes in amplitude and polarity will be obtained. The output signal or audio signal amplitude, as you have seen, is determined by the frequency deviation; for the less the frequency deviation, the less the departure from a 90-degree phase relationship between the reactive voltages E_2 and E_3, and E_1. The greater the frequency deviation, the greater is the difference in angular displacement between E_2 and E_1, and E_3 and E_1, so that the differential voltage obtained from the diodes is greater. When viewed from the angle of audio intensity, the greater the differential voltage from the rectifiers, the louder the audio signal, since the extent of deviation at the transmitter is a function of modulating voltage level. The

greater the modulating voltage level within prescribed limits, the greater the deviation frequency.

If the output voltage is plotted against input frequency, a response curve (identified as an "S" curve) is obtained. This is shown in Fig. 18. The perpendiculars A and B illustrate the positive and negative values of the output audio voltage for some arbitrary frequency deviations. The amplitudes of A and B do not represent the equal values of voltages applied across the diodes. These points (A and B) are representative of the extremes of deviation over which the output from the discriminator is linear.

The Effect of Amplitude Modulation

You will recall that the discusssion of discriminator action did not include any reference to amplitude-modulated signals appearing at the input to the discriminator itself. Everything discussed was in terms of frequency deviation in order to show how a differential output was produced. It so happens that the type of discriminator we described is also responsive to amplitude modulation of the input signal as well as frequency modulation. In
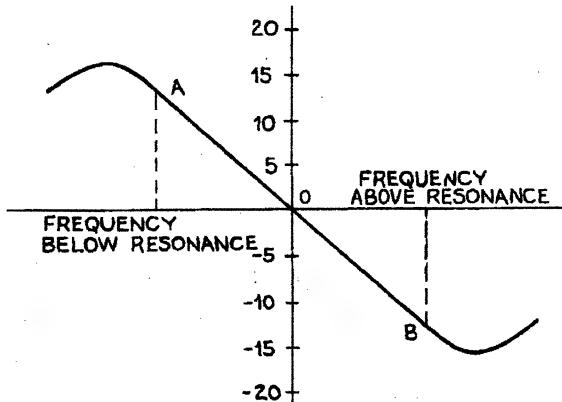

Fig. 18.—The "S" curve represents the typical output characteristic of the discriminator circuit with the output a-f voltage plotted against the input frequency deviations.

other words, if the limiter does not successfully clip the amplitude variations that may appear upon an f-m signal or if an amplitude-modulated signal finds its way into the input circuit of the discriminator, the discriminator will respond to these amplitude variations and consequently the output will contain amplitude-modulation effects. It can be readily seen that while uniform amplitudes of signals which

deviate equally from the carrier frequency, will appear as signals of like amplitude at the appropriate diodes, any variation in amplitude will cause a variation in the respective signal voltages.

Thus if for the sake of argument, at a signal of 4.325 mc the limiter output results in an amplitude of 2 volts and at a signal of 4.275 mc the limiter output results in an amplitude of 2.6 volts (these frequencies representing a 25-kc deviation from a center frequency of 4.3 mc), the respective voltages applied to the diodes will be different and the output will contain amplitude-modulation effects. Since the possible reasons for amplitude modulation are representative of undesired conditions, as noise or interference, it is highly desirable to eliminate all response to amplitude modulation in the discriminator.

The Ratio Detector

In the preceding paragraph it was shown how the conventional discriminator was in itself responsive to amplitude variations. This detection of a.m. was possible if the limiter did not function properly or if some a.m. happened to find its way into the input circuit of the discriminator. An f-m receiver not using a limiter stage but employing a detector that was unresponsive to a.m. but responsive to f.m., would be a decided advantage. Such a type of detector is known as the "ratio detector."

The ratio detector used in some f-m receivers is a discriminator because it converts frequency variations of the i-f carrier signal into amplitude variations, which appear as the

Fig. 19.—A typical ratio-detector circuit using a duo-diode tube. Note that the two diodes are connected in *series* through the resistor R. The a-f output voltage is taken off point D.

audio output from the f-m receiver. In addition to acting like a conventional discriminator, this type of detector is made unresponsive to any amplitude variations of the input signal to the detector; consequently, the use of limiter stages ahead of the detector is not necessary. As is the case in numerous basic circuits, various modifications of the ratio detector exist, and our discussion will center around one of these types.

All ratio detectors have two things in common; first, that phase shifting is the fundamental basis of converting frequency variations into amplitude variations (just as was described in the conventional discriminator circuit); second, the ratio detector output is proportional to the *changes in the ratio* between the two input signal voltages applied to the diodes. If you will recall, the conventional discriminator output represents the differential voltage developed across the diode loads and that no matter what the increase or decrease in output voltage the *ratio of the output voltage across the diode loads always remains the same* (for the same frequency), which output is proportional to the relative amplitudes of the input signal.

A typical ratio detector circuit showing the basic components, is illustrated in Fig. 19. Two diodes or a duo-diode tube can be used in the ratio-detector circuit similar to the conventional discriminator circuit, but in most cases a duo-diode tube is used. However, in the conventional discriminator circuit, the diodes are connected in an arrangement similar to push-pull, but in the ratio detector the diodes are connected in series. Examining Fig. 19, it will be seen that the plate of diode A is connected to the load resistor R, which in turn is connected to the cathode of diode B. The plate of diode B is connected to the cathode of diode A through the coil L_2 to complete the series path of these two diodes. This type of hookup affords a quick means of distinguishing between a ratio detector and a conventional discriminator circuit. The stage before the ratio detector is nothing more than the usual last i-f amplifier tube and it is often called the driver stage for the ratio detector. The load for this i-f tube is the tuned transformer, consisting of C_1, L_1 and C_2, L_2.

The voltage appearing across the tuned-primary circuit is coupled to the ratio-detector
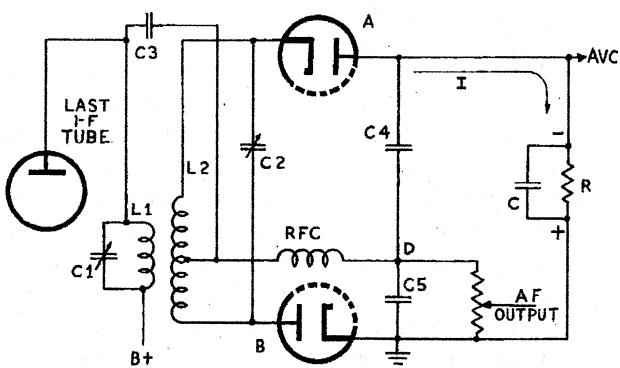
circuit in the two ways similarly described under the discussion of the conventional discriminator circuit: first, by transformer action from coil L_1 to coil L_2 and second, by direct coupling through C_3 to the r-f choke coil, RFC. Consequently, as was pointed out, the voltage appearing across L_1 likewise appears across the coil RFC in the same magnitude and phase relationship. Also equal but out-of-phase voltages appear across the individual halves of the coil L_2, the magnitude of these voltages depending upon the closeness of coupling between L_1 and L_2. The phase relationships among the voltages across coil L_1, the voltage across RFC, and the individual voltages across the separate halves of coil L_2, are the same as those discussed under the conventional discriminator circuit and illustrated by sine curves and vectorial pictures at resonance and off-resonance conditions to which the transformer is tuned.

Up to this point the analysis of the ratio detector and the previously described discriminator circuit is practically the same. From here on, however, the difference between the operation of these two f-m detectors manifests itself. Before understanding how this ratio detector removes amplitude modulation, the more fundamental operation of the rest of the circuit should be understood.

Examining the circuit of Fig. 19 once more, it is evident that since the diodes are connected in series they draw current in the same direction relative to R, which is also in series with them. Consequently, using the convention for the flow of electrons from cathode to plate, the current I will follow the path indicated by the arrow and the top part of resistor R will become negative with respect to its bottom or grounded end. If the two tuned circuits L_1, C_1 and L_2, C_2 are both resonant to the intermediate frequency and if an unmodulated i-f carrier signal (of the same frequency) were to be injected into the circuit from the i-f amplifier, then the two capacitors C_4 and C_5 will both be charged to the same voltage due to the symmetry of the circuit.

Now if the i-f carrier were frequency-modulated, then the voltages appearing across capacitors C_4 and C_5 would vary according to the modulation of the i-f carrier. The way this happens is as follows: It was mentioned that rectified current would flow in the direction shown in Fig. 19 and that the top portion of

the resistor R would have a negative potential on it. The values of the resistor R and capacitor C are so chosen that they represent a long time-constant network. Usually the value of this time-constant network can vary anywhere between one-tenth of a second to one-quarter of a second and still be effective to the desired degree. (The value of the resistance in megohms multiplied by the capacitance in microfarads will indicate directly the value of the time constant in seconds). Consequently, with a long time constant, it will take the capacitor C quite some time to discharge fully through R. Therefore the negative voltage at the top of resistor R will remain practically constant over the range of the lowest audio frequency desired to be reproduced in the output of the set. In other words, a time constant of one-tenth of a second corresponds to a frequency of 10 cycles per second; therefore, for frequencies above 10 cycles per second, the duration of one cycle would be shorter than the time constant and so the voltage across the R-C combination will remain practically constant. (The higher the audio frequency, the shorter the duration of one cycle.)

Since the voltage across R and C is constant, then the sum of the voltages across C_4 and C_5 must remain constant. However, if the carrier frequency falls below the i.f. or rises above the i.f., the voltages appearing across C_4 and C_5 will differ in value according to the degree of off-resonance conditions of the i-f signal. No matter what the difference between these voltages is, *their sum always remains the same.* Now if the i-f signal is frequency-modulated, the i.f. will vary above and below its resonant frequency according to the degree of frequency modulation. This accordingly will vary the voltages appearing across C_4 and C_5 but in a certain proportion determined by the potential across the RC combination. Consequently, it can then be said that the difference between the voltages across C_4 and C_5 appearing at the junction point D, varies at an audio rate (due to the degree of frequency-modulation.) Therefore, the a-f output is taken off between point D and ground and applied to the audio section of the set. According to this type of hookup the voltage appearing across C_4 will be larger than that across C_5 at frequencies below i.f., and above the i.f., the voltage across C_5 will be larger than that across C_4.

Removal of Amplitude Modulation

The basic part of the ratio detector in removing any amplitude modulation that appears in the circuit, is the R-C time constant network of Fig. 19. In other words it is the constant voltage across resistor R that plays the primary role in the removal of a.m. Let us suppose an a-m signal appears at the input of the ratio detector and see what happens:

Any a-m signal will tend to increase the output of the set and, of course, will increase the voltages across capacitors C_4 and C_5. However, the voltage across the RC network cannot change rapidly enough to follow the a.m. due to the nature of the long time constant, and the amplitude modulation therefore cannot change the voltage across C_4 and C_5. In other words, the capacitor C charges or discharges so slowly through R that the potential at the top of resistor R (or the plate of diode A) remains nearly constant and any amplitude modulation cannot change the voltage across capacitor C in step with this a.m. Consequently, sudden increases in amplitude of the f-m carrier will not have any effect in the output audio circuit because these sudden increases of amplitude cannot appear across either C_4 or C_5 as a change in voltage.

AVC From the Ratio Detector

In the usual f-m receivers employing a limiter tube with a conventional discriminator circuit, avc is taken off the resistor in the limiter control-grid circuit because of the negative voltage drop across this resistor due to grid current flow. In the sets employing a ratio detector, there is no limiter; however, the voltage across the resistor R in Fig. 19 serves as a means of obtaining avc voltage.

If capacitor C was removed from the circuit, then the voltage appearing across the resistor R will vary in accordance to the signal voltage of the received carrier. The capacitor C in conjunction with R offers a time-constant network that averages these changes appearing across R. That is, the time constant is not considered large in this instant as compared to the effect it has on removal of amplitude modulation, in that it permits slow changes of voltage across R and C in accordance with slow changes in the received signal voltage. Therefore, the

negative voltage at the top part of resistor R serves as a source of avc voltage.

Ratio Detectors Used In Today's Receivers

Since the ratio detector is a very recent development in f-m receivers, there are not many sets that use it. One circuit employing the ratio detector is illustrated in the Philco Model 46-480 Code 121, on Philco page 15-16 of Rider's Volume XV, in which a duo-diode 6H6GT/G tube is used in the ratio-detector circuit, preceded by a 7H7 second i-f amplifier. The circuit may appear somewhat different than that illustrated in Fig. 19, but basically the underlying action is the same.

In Fig. 20 is shown a ratio detector circuit as used in some of the RCA Victor f-m receivers that employ ratio detectors. The fundamental action of this circuit is the same as that of Fig. 19.

Alignment of Ratio Detectors

The ratio detector of Fig. 20 used in RCA Victor receivers is aligned as follows:

1. Short L_1 with a short piece of wire.

2. With an unmodulated carrier tuned exactly to the receiver i.f. applied to the driver grid, peak the driver tank for maximum avc voltage as measured with a high-resistance voltmeter.

3. Remove the short on L_1 and connect two matched 75,000-ohm resistors in series across R_5.

4. Connect a high-resistance voltmeter between the junction of the two matched resistors and the junction of C_3 and C_4.

5. Using an amplitude-modulated carrier (30% at 400 cycles) tuned exactly to the receiver i.f., and fed in at the driver grid, adjust the slug marked "tune" for zero balance on the voltmeter, and the slug marked "center" for minimum audio output at the speaker.

6. Repeat step 2 without the short on L_1.

7. Repeat step 5 making sure that the "tune" and "center" adjustments occur simultaneously.

In aligning ratio detectors, the zero-balance adjustment is the adjustment obtained by tuning from a positive voltage through zero to a negative voltage or vice versa, not the zero voltage reading obtained when the transformer is tuned to a frequency far removed from the
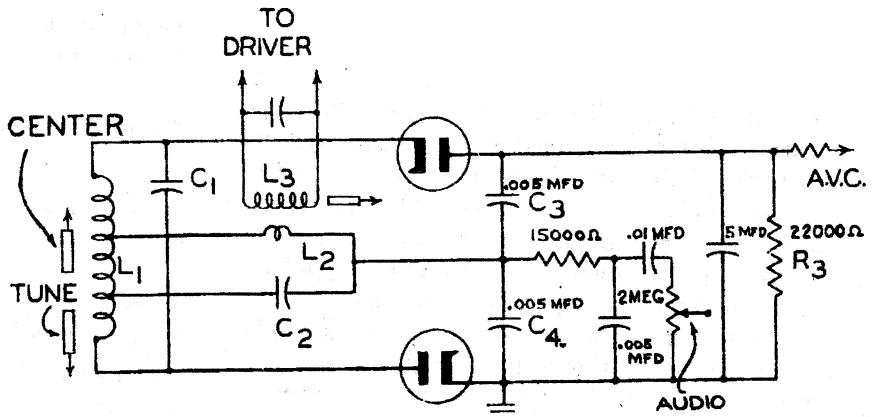
Fig. 20.—The ratio-detector circuit employed by RCA in several of their f-m receivers, the first of which is Model 612V1.

Courtesy RCA Mfg. Co.

nominal i.f. of the receiver. Once the true zero balance is obtained, it will be possible to obtain a negative voltage by slightly shifting the balance adjustment in one direction, and a positive voltage by shifting the same adjustment in the other direction. If a vacuum-tube voltmeter is used for the zero balance adjustment, the serviceman should be sure that the meter does not ground the junction of the two 75,000-ohm resistors through the power line. A Volt-Ohmyst can be used by connecting the "COMMON" lead to the junction of the two 75,000-ohm resistors, and the "D-C VOLTS" lead to the junction of C_3 and C_4, making sure that the Volt-Ohmyst chassis does not come into contact with the receiver chassis.

When aligning some ratio detectors, it may be necessary to disconnect the driver grid from its resonant grid circuit, and re-connect the grid to ground through a 50,000-ohm resistor. The signal generator output will then be fed directly into the driver grid.